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LONG FEEDERS FOR TRANSMITTING WIDE SIDE-BANDS, WITH REFERENCE TO THE ALEXANDRA PALACE AERIAL-FEEDER SYSTEM*

By E. C. CORK, B.Sc.(Eng.), Associate Member, and J. L. PAWSEY, Ph.D.†

(*Paper first received 7th March, and in revised form 22nd October,* 1938; *read before the* Wireless Section *7th December,* 1938.)

SUMMARY

The paper is concerned with reflection phenomena in long feeders used to transmit modulated high-frequency carrier waves such as television signals, in which the time of transmission of a wave along the feeder is comparable with the time periods of the modulation frequencies.

The nature of the various impedance irregularities giving rise to reflection is discussed, and their effect on a received television picture is indicated. Methods for reducing the effect of these irregularities are considered.

These reflecting irregularities may be studied in terms of the variation with frequency of the input impedance of the terminated feeder, and examples are given of the magnitude of the effects for various feeders.

The vision aerial feeder system of the London Television Station is described, including the aerial-to-feeder matching arrangements and a device for reducing the residual mismatch due to the variation of the aerial impedance over the frequency range of the carrier and side-bands.

Apparatus for measuring the input impedance and the methods adopted for establishing a standard resistance for terminating the feeder are described.

In the Appendix consideration is given to the permissible limits of eccentricity of an approximately terminated but irregularly eccentric feeder.

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(1) INTRODUCTION

This paper is concerned with the problems which arose in the course of the design of the aerial-feeder system for the Alexandra Palace Television Station erected in August, 1936.

In the preceding year complete vision and sound aerial and feeder systems were erected on a mast similar to that at Alexandra Palace at the E.M.I. Laboratories at Hayes for experimental purposes. In the course of these investigations it became apparent that there were stringent impedance-matching conditions to be satisfied if distortion of the transmitted picture was to be avoided. These conditions are connected with the fact that the time of travel of a wave along the feeder is not negligible compared with the time periods of the modulation frequencies. These conditions appear to have hitherto received little attention. The special requirements and the methods adopted to fulfil them are discussed in some detail.

Apparatus was designed to measure impedance at 45 Mc./sec. to an accuracy of a few per cent, and such measurements were fundamental to the design of transformers and impedance-correcting devices employed. The apparatus was further utilized to check the characteristics of the system at various stages of its construction.

At an early stage of the work it was necessary to produce a feeder-terminating resistance of the value of the characteristic impedance of the feeder which should be constant and non-reactive over a frequency range of from 43 to 47 Mc./sec. The difficulty in doing this will be appreciated when it is pointed out that at these frequencies the reactance of a lead consisting of an inch of wire may be of the order of 5 ohms.

The methods of measurement are described and curves are given which show the impedance of the aerial feeder system over the frequency range 42 to 48 Mc./sec. at important stages in the construction of the system.

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The Alexandra Palace Television Station was designed to operate on a vision frequency of 45 Mc./sec. and a sound frequency of 41·5 Mc./sec. The pictures to be transmitted were 405 lines, 25 pictures per sec. with interlaced scanning, giving 50 frames per sec. This requires side-bands of approximately 2·5 Mc./sec.

The length of the feeder from the transmitter to the aerial is approximately 450 ft., so that the time of travel along it is 0·5 μsec., corresponding to a frequency in the side-band range.

(2) DISTORTION DUE TO REFLECTIONS IN THE FEEDER

A form of distortion of the transmitted picture with which this paper is particularly concerned is caused by reflected waves in the feeder due to the mismatch between the aerial and the feeder, and the transmitter and the feeder.

In order to obtain the wide side-bands required in the Alexandra Palace installation, it was necessary to arrange that the final output circuit of the transmitter should be heavily damped. The damping is due to the losses in the valves and circuits, together with the power supplied to the aerial. The damping due to the losses is small, and the major part is supplied by the aerial feeder system. This involves a large mismatch at the transmitter end of the feeder.

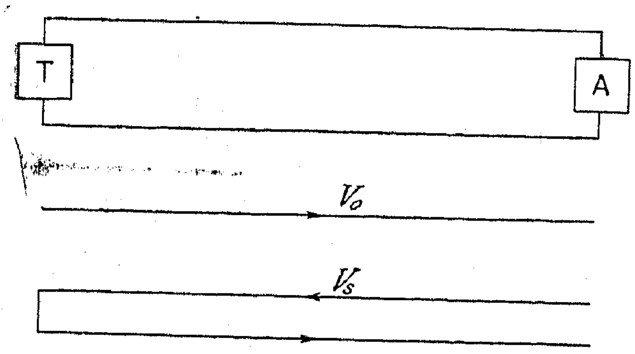

Fig. 1

On account of the extremely low loss of the feeder and the high degree of mismatch at the transmitter end, a wave, if reflected from the aerial end, passes to the transmitter and arrives again at the aerial scarcely diminished in amplitude, but delayed by twice the time of travel along the feeder. This time-delay is about 1 μsec., during which time the scanning spot on a receiver screen will have travelled 1/100th of the width of the screen, a distance equal to that occupied by about 5 lines in the vertical direction. If the picture contains a thin vertical line it is obvious that reflections will give rise to distortion consisting of "echo" lines displaced from the original image at intervals of 1/100th of the picture width. A sharp edge would be followed by similar striae. Various forms of this distortion were observed in the early experiments at Hayes.

In order to estimate the accuracy of the termination required to avoid objectionable distortion, consider the transmission of a picture containing an abrupt transition from white to black, involving a sudden transition from full carrier to, say, zero carrier. If the final value is not zero the discussion is complicated by interference between the existing steady wave and the distorting waves to be described, but is otherwise unmodified.

Let the transmitter T (see Fig. 1) be of high impedance, so that substantially perfect reflection takes place at T of any wave arriving from the direction of the aerial A. During the transmission of the white portion of the picture the aerial current will be due to the resultant of the direct wave, of voltage V_0, initiated by the transmitter, and, if the aerial does not exactly terminate the feeder, a steady reflected wave V_s reflected from A and T in succession. Higher-order reflected waves will be neglected, since they must be of smaller magnitude than the first, and it will be shown that the first must be made negligible to avoid distortion. If l is the length of the feeder and c the velocity of the wave in the feeder, then at a time l/c after the stoppage of the transmitter the voltage on the aerial due to V_0 falls to zero, but that due to V_s persists a further time $2l/c$. This gives rise to a "step" in the decay of aerial current as indicated in Fig. 2. The relative height of the "step," if V_s is small, is given by the reflection coefficient $(Z_a - Z_0)/(Z_a + Z_0)$, where Z_a and Z_0 are respectively the terminating and characteristic impedance of the feeder.

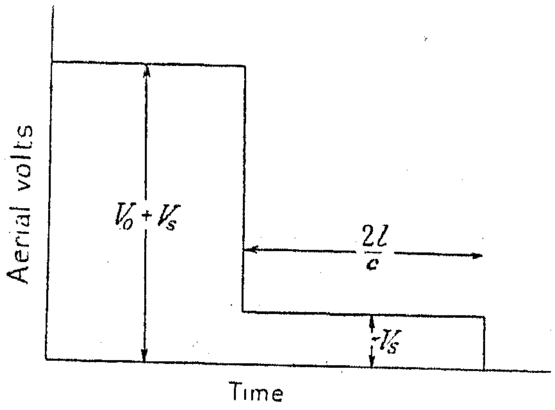

Fig. 2

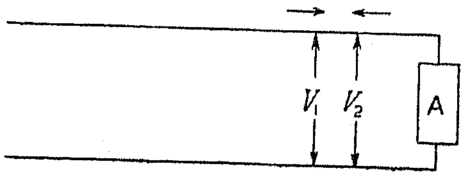

Fig. 3(*a*)

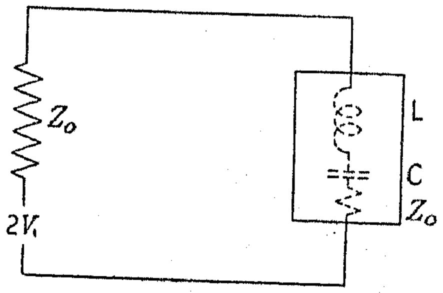

Fig. 3(*b*)

If the aerial is a resonant system then a further important factor must be taken into account. On the sudden decay of impressed volts from $(V_0 + V_s)$ to V_s the aerial current will not fall instantaneously to the new steady state but will decay with a period and damping determined by the constants of the aerial. This effect would be much less important but for the fact that during the decay the aerial transfers part of its stored energy to the feeder, originating a wave which arrives back at the aerial after a delay of $2l/c$ and causes a further short burst of aerial current.

The magnitudes involved follow from the following considerations. If V_1 represents the voltage of the forward wave at the aerial end of the feeder (see Fig. 3*a*), and V_2 the reverse wave, then the voltage across the termination V_a equals $(V_1 + V_2)$. Further, the current and voltage across A due to the incident wave V_1 are equal to those which would be caused by an e.m.f. $2V_1$ in series with Z_0 and the aerial, as indicated in Fig. 3(*b*).

On the subsequent return of this wave, for which now $2V_1 = 2|V_0|e^{-\alpha t}\cos\omega_0(t - 2l/c)$, the transient aerial current i_t, calculated from the circuit of Fig. 3(*b*), is given,* neglecting the time-delay term, by

$$i_t = \frac{|V_0|te^{-\alpha t}}{L}\cos(\omega_0 t + \Psi) \quad . \quad . \quad . \quad (2)$$

The envelope is given by

$$\frac{|V_0|}{L}te^{-\alpha t} = \frac{|V_0|}{Z_0}\alpha te^{-\alpha t} = i_a\alpha te^{-\alpha t} \quad . \quad . \quad (3)$$

where i_a is the magnitude of the steady aerial current due to V_0.

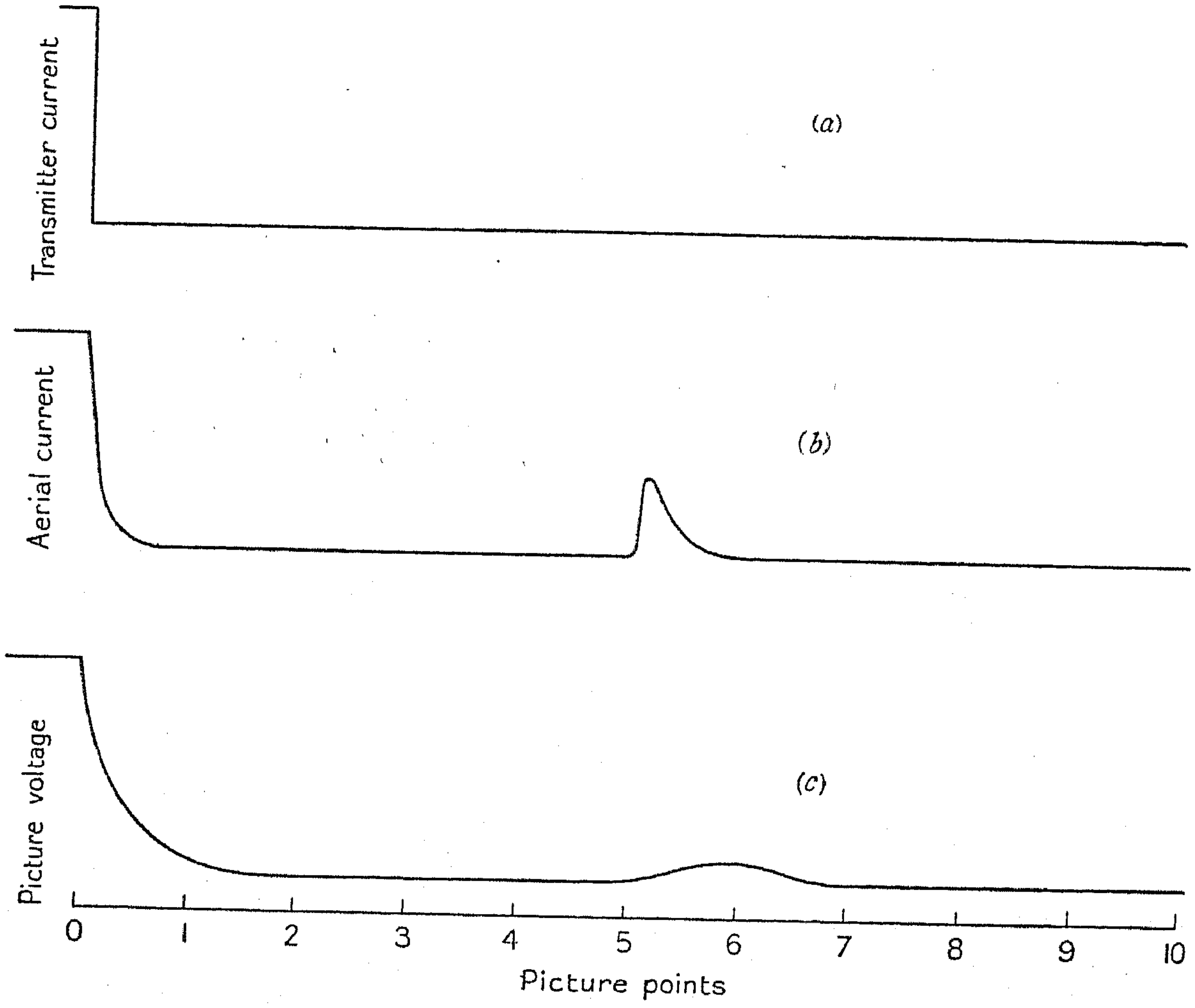

Fig. 4

(*a*) Transmitter current (envelope).
(*b*) Aerial current (envelope).
(*c*) Receiver picture wave-form.

As an illustration, consider A to be an inductance L in series with a capacitance C, tuned to the carrier frequency, and a resistance Z_0, so that there is no steady reflected wave V_s. Let the incident wave $V_1 = |V_0| \cos\omega t$ fall to zero at the aerial at time $t = 0$. The transient voltage across the aerial will be of the form

$$V_a = Ae^{-\alpha t}\cos(\omega_0 t + \phi) \quad . \quad . \quad . \quad (1)$$

where $\omega_0 = \sqrt{\left[\frac{1}{LC} - \left(\frac{2Z_0}{2L}\right)^2\right]}$

and $\alpha = \frac{2Z_0}{2L}$.

Also, from the initial conditions, $A = V_0$ and $\phi = 0$. Further, since after time $t = 0$, $V_1 = 0$, then $V_2 = V_a$ and the whole of this voltage is reflected.

The term $\alpha te^{-\alpha t}$, which determines the relative amplitude of the envelope of the burst of aerial current, reaches a maximum amplitude of $1/e$ when $t = 1/\alpha$, and then decays. The amplitude of this burst of aerial current is quite high and, moreover, not controllable. The duration, however, is very short and is a function of the aerial decrement α. Fig. 4 shows, in succession, (*a*) the envelope of the postulated transmitter current, and (*b*) the envelope of the aerial current as given by equation (3) for a half-wave dipole of $\frac{1}{16}$ in. diameter wire, tuned and matched at the carrier frequency and fed over a feeder 450 ft. in length. Such an aerial has a decrement α of about 40×10^6/sec.

The delayed burst of aerial current when acting on a receiver would tend to produce a picture voltage of reduced relative amplitude and extended duration on account of the selective circuits of the receiver. The effects would be the smaller the less the duration of the impulse. Fig. 4(*c*) shows in a qualitative manner the received picture wave-form on an idealized receiver. An estimate by Fourier methods of the amplitude of the disturbance in the picture wave-form due to the delayed burst of aerial current of Fig. 4(*b*), assuming an idealized receiver with a pass band of ± 2 Mc./sec., gave a value of about 10 % of the preceding steady voltage and indicated that for further increases in the decrement the amplitude would vary inversely as the decrement.

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* See V. Bjerknes: *Wiedemann's Annalen*, 1895, vol. 55, p. 121.

Though this analysis has considered the special case of a simple tuned circuit, a similar argument would appear to apply to other forms of resonant aerials.

To reduce the magnitude of the effect the aerial may be made to have a sufficiently high decrement by some means such as the employment of thick conductors, and also a certain amount of correction may be performed by inserting additional circuits which contribute transients in anti-phase to the original transient. A simple form of this latter arrangement in the case of the aerial circuit discussed would consist of a series-tuned circuit of the same L and C as the aerial, inserted in series with the feeder at one quarter-wavelength from the aerial. The resulting waves reflected down the feeder are indicated in Fig. 5, and it is apparent that the second circuit may effect a substantial degree of cancellation. It is obviously desirable that the two points of reflection in the feeder should be as close together as possible in order that as few half-cycles as possible may remain uncancelled. Beyond the stage at which the effect of the first transient is over before the arrival of the second no cancellation can be accomplished, but this limit to the distance which may separate the points of reflection is fixed not by the duration of the aerial transient itself, but by the duration of the effect on the final receiver circuit, since two completely isolated transients incident in succession on the receiver in appropriate phases may produce a resultant less than that due to either alone.

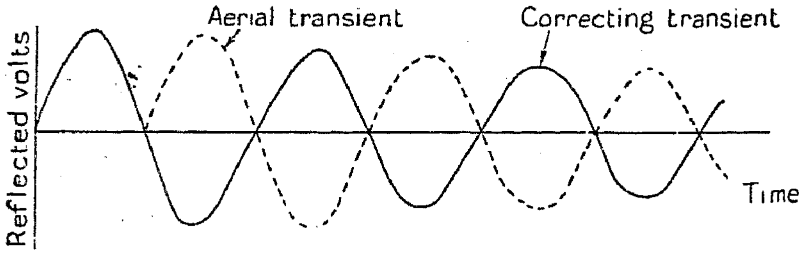

Fig. 5

An alternative method of considering this phenomenon is in terms of the Fourier components of the signal. If the feeder is misterminated at any frequency within the range of the carrier and side-bands, the input impedance will vary from the characteristic impedance over this frequency range. Such a varying impedance is accompanied by a varying transfer of the Fourier components of the signal from the transmitter to the feeder. On account of the large mismatch at the transmitter, the energy transferred at one frequency will vary approximately linearly with the input impedance, and, since the loss in the feeder is negligible, the same energy is radiated from the aerial. These considerations indicate that it is a necessary condition for the avoidance of distortion that the input impedance to the feeder should be constant over the frequency range to be transmitted. This implies that the aerial should match the feeder over the whole of the frequency range, and, in addition, all causes of reflection in the feeder itself must be reduced to low values.

In order to estimate the tolerance which may be allowed in fulfilling these conditions without introducing appreciable distortion in the transmitted pictures, it is necessary to assign a value to the minimum change in signal which produces a discernible change of intensity in the received picture. This change of signal is dependent on the existing carrier level, and is very small in the neighbourhood of 30 % of the maximum, which in the Marconi-E.M.I. system is the picture-black level. In this region a change of carrier of the order of 2 % of the maximum can be observed.

Accepting this figure, it follows from the preceding discussion on a sudden transition in picture voltage from full white to black that the aerial must match the feeder at the carrier frequency to an accuracy of at least 2 × 2 % to avoid a visible step in the picture.

The permissible limit to the mismatch at the side-band frequencies is difficult to assign. In the case discussed of a resonant aerial matched at the carrier frequency (see Fig. 4) the distortion was 5 times that to be allowed. It would therefore appear that a decrement of not less than 5 times that of the case of Fig. 4, i.e. 200×10^6/sec., is desirable. A simple tuned circuit with this decrement would be mismatched by 12 % when 2 Mc./sec. off tune, and this figure is suggested as an estimate of the greatest unobjectionable mismatch throughout the frequency range 43 to 47 Mc./sec.

(3) FORMULAE FOR THE INPUT IMPEDANCE OF APPROXIMATELY MATCHED FEEDERS

(a) Localized Impedance Irregularities

In the course of the investigations the following approximate transmission-line theory was found very useful in reducing the labour of calculations and giving a clear picture of the processes involved.

It can be readily shown that the input impedance Z_i of a feeder of characteristic impedance Z_0 and of electrical length θ (Fig. 6), when terminated by an impedance $(Z_0 + a)$, is approximately given by

$$Z_i = Z_0 + ae^{-2\theta} \quad . \quad . \quad . \quad . \quad (4)$$

This form is exact for small values of a and is accurate to 0·05 % of Z_0 for $|a|/Z_0 < 0 \cdot 1$. In the case of a feeder of negligible loss, such as the Alexandra Palace feeder, θ may be replaced by $j\alpha$ and the formula becomes

$$\begin{aligned} Z_i &= Z_0 + ae^{-2j\alpha} \\ &= Z_0 + |a|e^{j(\phi - 2\alpha)} \end{aligned}$$

where a is the complex quantity $|a|\underline{/\phi}$ and so may be written $|a|e^{j\phi}$. For an essentially air-spaced feeder $\alpha = 2\pi lf/c$, where l is the length of the feeder, c the velocity of light, and f the frequency. Now the quantity

$e^{-2j\alpha}$, which equals $\cos 2\alpha - j \sin 2\alpha$, is of unit magnitude and of phase angle -2α. Equation (4) may therefore be interpreted as follows. The input impedance is equal to the sum of the characteristic impedance Z_0 and the deviation of the terminating impedance from Z_0, the latter term being rotated through an angle -2α. This rotation through an angle -2α is related to the wave reflected from the mistermination having to travel twice the length of the feeder in passing from the input to the termination and then back again.

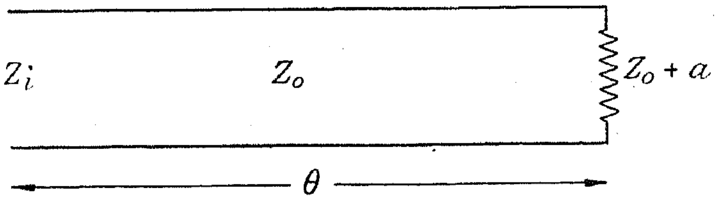

Fig. 6

The equation may be represented vectorially for any given values of Z_0, a, and α, as indicated in Fig. 7. Here OA represents the characteristic impedance and $\text{OB} = (Z_0 + a)$ is the terminating impedance. The input impedance OC is found by rotating the vector AB through the angle -2α to AC. The input resistance r_i and reactance x_i are given by OD and DC respectively, where D is the foot of the perpendicular from C on the line OA. The corresponding algebraic formulae are

$$\left.\begin{aligned} r_i &= Z_0 + a\cos(\phi - 2\alpha) \\ x_i &= a\sin(\phi - 2\alpha) \end{aligned}\right\} \quad . \quad . \quad . \quad (5)$$

Since $\alpha = 2\pi l f/c$ it follows that the input impedance is a function of the feeder length and frequency. For a fixed value of a the deviation of the input resistance from Z_0 has a sinusoidal variation with frequency of " periodic interval " equal to $c/(2l)$, and the reactance a similar variation in quadrature with the first. The word " periodic interval " is here used to denote the interval in the variable, in this case the frequency, between the successive points at which the function is in the same phase. In practice it is sufficient to measure the variation of resistance with frequency, since reactance provides no additional information.

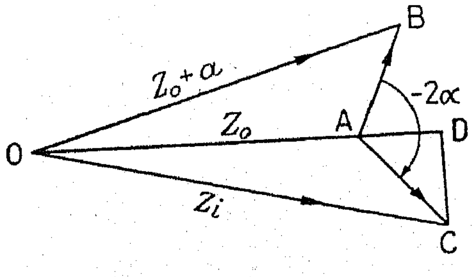

Fig. 7

Referring to Fig. 8 and applying equation (4) in succession to the points P and Q, it can be readily shown that the input impedance of a length of feeder terminated with the impedance $(Z_0 + a_1)$ and having an additional localized series impedance a_2 at a distance α_2, is given by

$$Z_i = Z_0 + a_1 e^{-2j\alpha_1} + a_2 e^{-2j\alpha_2} \quad . \quad . \quad (6)$$

Similarly for any number of localized series impedance irregularities the individual terms of the form $ae^{-2j\alpha}$ are additive, corresponding to addition of the individual reflected waves. The variation of the input resistance or reactance with frequency of a feeder with a number of fixed impedance irregularities is therefore given by the sum of a number of simple harmonic terms each of an amplitude dependent on the magnitude of the irregularity and of a " periodic interval " dependent on its distance from the input end of the feeder.

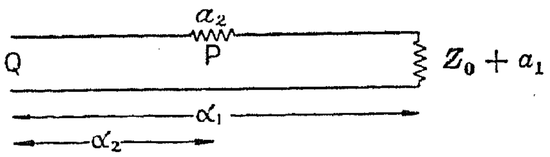

Fig. 8

The discussion so far has been restricted to impedance irregularities which are in series with the feeder, such, for example, as could be caused by an imperfect contact at the joint between two sections. A common type of irregularity is an irregularity in the impedance between the two conductors, such as a localized decrease of impedance due to the presence of an insulator. Such cases are more readily treated in terms of admittances than of impedances. The basic formula for the input admittance Y_i, of a feeder of electrical length θ and of characteristic admittance Y_0, defined as $1/Z_0$, when terminated by an admittance $(Y_0 + A)$, is

$$Y_i = Y_0 + Ae^{-2\theta}$$

which is of the same form and subject to the same limitations as the impedance equation, equation (4). It follows that the derived equations for impedances and admittances are of similar forms. The latter type of irregularity may therefore be treated in terms of the admittance irregularity between the conductors of the feeder, or, alternatively, it is readily possible to transform the admittances to equivalent series impedances.

(b) Extended Impedance Irregularities

The effect of inserting a section of feeder of different characteristic impedance may be obtained by considering the input impedance of the feeder shown in Fig. 9 at the points N, L, M, successively. By this means it can readily be shown that the input impedance at M is approximately given by:—

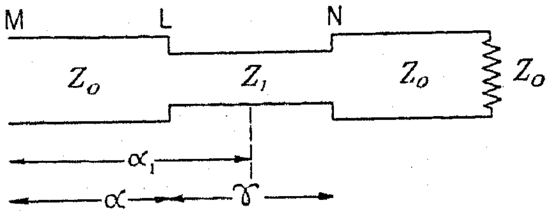

Fig. 9

$$\begin{aligned} Z_i &= Z_0 + 2j\sin\gamma\,(Z_1 - Z_0)e^{-2j(\alpha + \frac{1}{2}\gamma)} \\ &= Z_0 + 2j\sin\gamma\,(Z_1 - Z_0)e^{-2j\alpha_1} \quad . \quad . \quad (7) \end{aligned}$$

where $\alpha_1 = \alpha + \frac{1}{2}\gamma$. This equation is identical with that applicable to a localized impedance irregularity a inserted at the mid-point of the section of the feeder of abnormal characteristic impedance, provided that

$$a = 2j(Z_1 - Z_0)\sin\gamma$$

The irregularity gives rise to fluctuations of input impedance similar to those already discussed.

It is possible to reduce the effect of such a section of feeder either by making it short or by making it approximately a multiple half-wavelength in length so that $\sin\gamma = 0$. Alternatively, the effect may be cancelled by the insertion of a localized impedance irregularity such as a condenser across the feeder or a similar section of feeder an odd multiple of a quarter-wavelength from the centre of the abnormal section. Such a cancellation may hold sufficiently over the necessary frequency range if the spacing of the opposing irregularities is small.

(4) IMPEDANCE IRREGULARITIES IN THE FEEDER

(a) Insulators and other Capacitive Irregularities

In a completed aerial and feeder system the impedance at the transmitter is a function of the aerial impedance, together with any impedance irregularities which may occur along the length. Mechanical considerations require the provision of bends, joints, insulators, etc., any one of which may cause an impedance irregularity of appreciable magnitude.

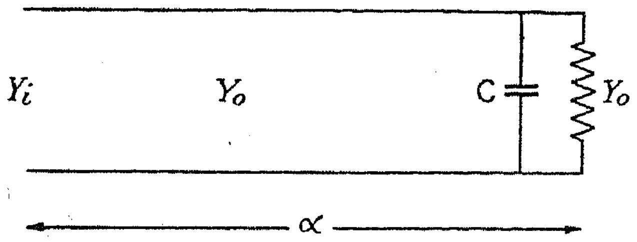

Fig. 10

Consider a single insulator in an otherwise uniform and correctly terminated feeder as shown in Fig. 10. If the insulator is substantially free from loss the admittance A introduced by its presence is equal to $j\omega C$, where C is the capacitance introduced and the input admittance is given by

$$Y_i = Y_0 + j\omega C e^{-2j\alpha}$$

and the conductance by

$$G_i = Y_0 + \omega C \sin\frac{4\pi lf}{c} \quad . \quad . \quad . \quad (8)$$

The insulator therefore produces a sinusoidal variation of G_i with frequency. In the case under consideration ωC is much less than Y_0, and the inverse of the conductance, i.e. the parallel resistance R_i, is given very nearly by

$$R_i = Z_0 - \omega C Z_0^2 \sin\frac{4\pi lf}{c} \quad . \quad . \quad . \quad (9)$$

The magnitude of C for various insulators considered ranged from 0·2 to 2 $\mu\mu$F, giving at 45 Mc./sec., for the type of feeder used, a fluctuation in R_i of from 0·5 % to 5 %, so that the effect of even a single insulator was large.

Equation (9) is important because it enables the magnitude and location of a capacitive impedance irregularity to be determined from measurements of the input resistance of the terminated feeder over a frequency range. Fig. 11 illustrates a possible curve of input resistance against frequency. Since it is sinusoidal it may be inferred that the impedance irregularities are restricted to a small length of feeder and so are probably due to a single cause. The "periodic interval" of the oscillation $(f_1 - f_2)$ locates the irregularity at a distance of $c/[2(f_2 - f_1)]$ from the input end, and an inspection of the feeder in this vicinity should show the cause. If this cause is assumed to be a local excess capacitance, which is common, it is possible to locate it accurately without precise frequency measurement by means of the phase of the resistance oscillation. Inspection of equation (9) shows that the frequencies at which R_i increases through the mean value are those for which the electrical lengths of the feeder to the point are odd multiples of a quarter-wavelength. Let the feeder be short-circuited in the vicinity of the capacitive irregularity, and the frequencies observed at which the short-circuited feeder is an odd multiple of a quarter-wavelength. If these frequencies are F_1 and F_2, etc. (see Fig. 11), then the electrical length to the short-circuited point is less than that to the irregularity by

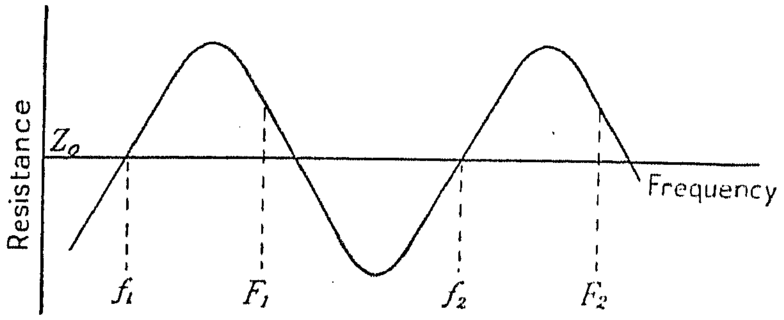

Fig. 11

$$\left(\frac{F_1 - f_1}{f_2 - f_1}\right)\frac{\lambda}{2} + \frac{n\lambda}{2}$$

where n is any integer. It is assumed that the measurement of the "periodic interval" is sufficiently accurate to determine n. Since it may be assumed that the velocity down the feeder is that of light, the physical position of the irregularity is thus located.

If it is not possible to remove the impedance irregularity, the effect may be substantially annulled by the insertion of an additional condenser of equivalent magnitude one quarter-wavelength distant from the point determined above. This method of cancellation was used when necessary at Alexandra Palace to reduce the variations in somewhat more complex curves than that of Fig. 11. The positions and magnitudes of the correcting condensers were determined by superposing curves such as that of Fig. 11 on the experimental curve and choosing an amplitude, frequency, and phase, to best effect cancellation. The type of condenser used consisted of a disc of brass on the end of a brass spindle screwed through the outer wall of the feeder. This disc introduced an excess capacitance between inner and outer, variable between about 0·2 and 1 $\mu\mu$F.

In the case of two insulators in an otherwise uniform feeder, there is a variation of the input conductance which is the sum of two sinusoidal terms; thus

$$G_i = Y_0 + \omega C_1 \sin\frac{4\pi l_1 f}{c} + \omega C_2 \sin\frac{4\pi l_2 f}{c} \quad . \quad (10)$$

In the particular case of two similar insulators, for which $C_1 = C_2$, the deviations from Y_0 cancel if $2\alpha_1 - 2\alpha_2 = \pi$, 3π, 5π, etc., i.e. if the insulators are separated by an odd multiple of a quarter-wavelength. Further, if the frequency changes over a small range the departure from this condition for a small spacing is not great and the cancellation is substantially maintained.

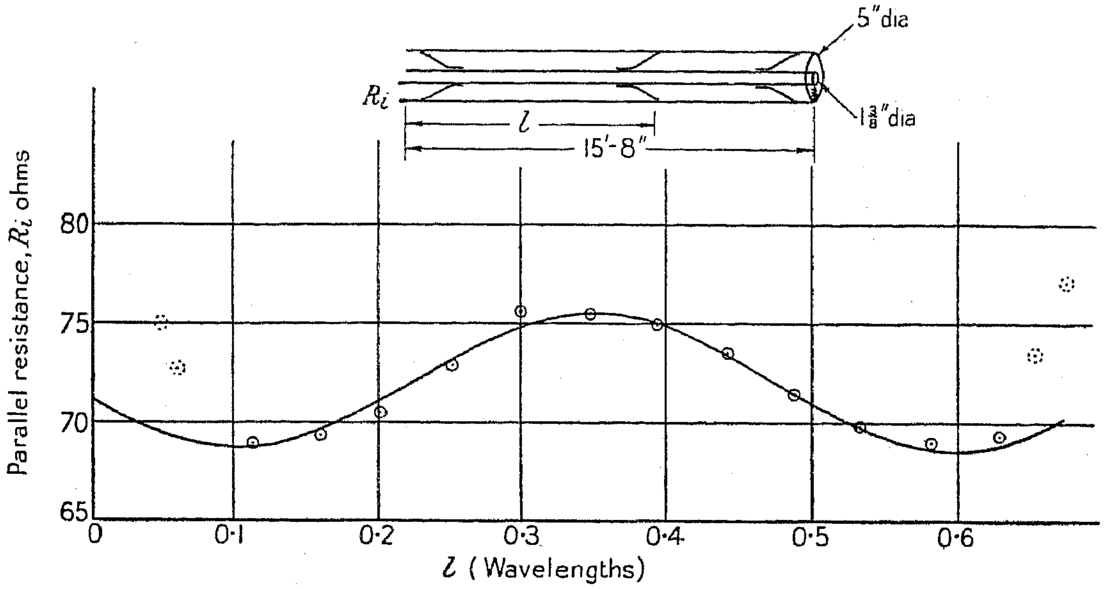

Fig. 12

In general the variation of G_i with frequency indicated in equation (10) is that of the sum of two simple harmonic terms of "periodic interval" $c/(2l_1)$ and $c/(2l_2)$, respectively. This gives rise to a combined oscillation of fluctuating amplitude which is analogous to the well known "beat" phenomena in sound. The maximum amplitude when the two are in phase is $(\omega C_1 + \omega C_2)$, and the minimum $(\omega C_1 - \omega C_2)$. The "periodic interval" of the combined oscillation is $c/(l_1 + l_2)$, and the interval between successive beat maxima is $c/[2(l_1 - l_2)]$, corresponding to the mean and the difference of the individual frequencies comprising the oscillations. In the same manner the susceptance shows similar phenomena displaced 90° in phase.

These forms of interference can, of course, occur equally well in the case of other forms of irregularities.

(b) The Experiment of Drawing Insulators Along a Concentric Feeder

To confirm the preceding theory and to obtain the order of magnitude of the resistance variations due to insulators, in the preliminary investigations experiments were made in which insulators were pulled along a length of feeder by means of a piece of string. The feeder was terminated at the far end with a resistance approximately equal to its characteristic impedance, and the input resistance was measured for various positions of the insulator. Fig. 12 shows the general arrangement, comprising a movable bell-type porcelain insulator between two fixed ones at the ends of the feeder. The input parallel resistance is shown in the graph plotted against l measured in wavelengths. The variation is approximately sinusoidal except near the ends of the feeder, in which positions the inner conductor sagged, so vitiating the results in these regions. The "periodic interval" of the resistance variation is approximately a half-wavelength, and the amplitude of the variation, 3·7 ohms, corresponds to a capacitance of 2 $\mu\mu$F. Low-frequency measurements (1 000 cycles per sec.) of the capacitance of a section of feeder with and without the insulator in position gave the same value. The mean value of the resistance variation is approximately the characteristic impedance of the feeder, but is modified by a number of constant factors such as the fixed insulators and the accuracy of the termination.

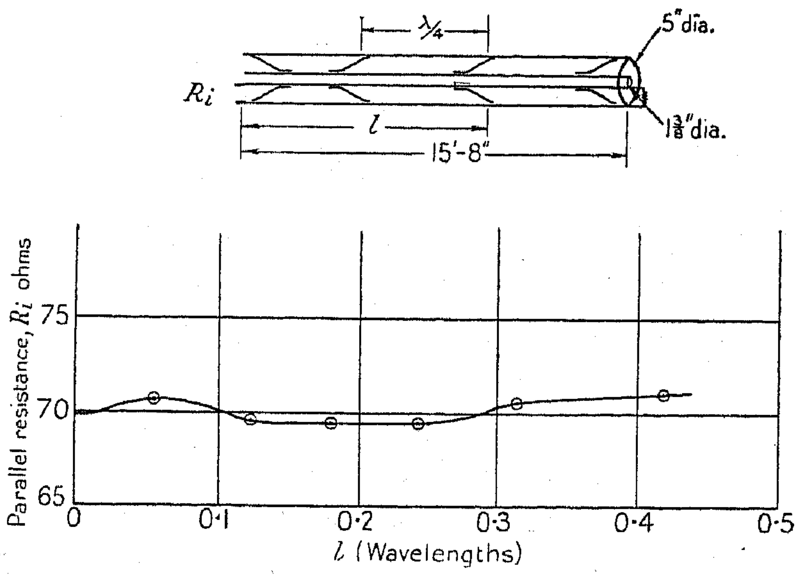

Fig. 13

A similar experiment was performed to investigate the cancelling effect of two insulators at a quarter-wavelength spacing. The arrangement is shown, together with the measurements of the resistance, in Fig. 13. The curve shows that the cancellation is complete to 0·5 ohm, which was about the accuracy of the measurement in these preliminary experiments.

A variation of this method was adopted for measuring the capacitance introduced by the various types of insulators. A comparatively short length of terminated feeder was used and the input resistance measured. The feeder was then taken to pieces, one or more insulators were inserted in the position corresponding to the first minimum of Fig. 12, and the feeder was then re-assembled. The change of the input resistance was then a measure of the capacitance change.

(c) Uniform Spacing of Insulators

The preceding Section has indicated that even a single insulator may give rise to large variations of input impedance, but that it is possible that insulators may be so arranged that their effects cancel. A particular case of considerable importance is that of a feeder having uniformly spaced insulators. The previous analysis neglected the second-order terms due to the interaction between the insulators, and its validity is doubtful for a large number of insulators when the sum of the first-order terms is nearly zero.

An exact analysis of this case may be obtained by considering this loaded feeder replaced by a new feeder of modified constants to which it is equivalent at a given frequency. The loaded feeder may be broken up into π sections at the middle of each insulator, the admittance of the insulators added to the parallel members, and the constants of the equivalent feeder sections obtained. In this way it may be shown that if:—

Z_0 = characteristic impedance of the feeder without insulators,
Z_0' = characteristic impedance of the feeder with insulators,
l = insulator spacing,
λ = wavelength in the unloaded feeder,
λ_1 = wavelength in the loaded feeder,
θ = electrical length of a section of unloaded feeder,
θ_1 = electrical length of a section of loaded feeder,
$\alpha = \dfrac{2\pi l}{\lambda}$.
$\alpha_1 = \dfrac{2\pi l}{\lambda_1}$,
C = capacitance of each insulator,
$\dfrac{1}{Z_a}$ = admittance change due to the presence of an insulator,

then

$$Z_0' = \frac{Z_0}{\left[1 + \left(\frac{Z_0}{2Z_a}\right)^2 + \frac{Z_0}{Z_a}\coth\theta\right]^{\frac{1}{2}}} \quad . \quad (12)$$

$$\sinh\frac{\theta_1}{2} = \sinh\frac{\theta}{2}\left(1 + \frac{Z_0}{2Z_a}\coth\frac{\theta}{2}\right)^{\frac{1}{2}} \quad . \quad . \quad (13)$$

It may be assumed that the losses in the feeder and insulators are negligible, so that

$$\theta = j\alpha \quad \text{and} \quad \theta_1 = j\alpha_1, Z_a = \frac{-j}{C\omega}$$

and also that $\left(\dfrac{Z_0}{2Z_a}\right)^2$ is negligible, so that

$$Z_0' = \frac{Z_0}{(1 + Z_0C\omega\cot\alpha)^{\frac{1}{2}}} \quad . \quad . \quad . \quad (14)$$

and

$$\cos\alpha_1 = \cos\alpha - \tfrac{1}{2}Z_0C\omega\sin\alpha \quad . \quad . \quad (15)$$

It is evident that the constants of the equivalent feeder are a function of frequency, and hence over a range it would require a terminating resistance varying with frequency in order to avoid reflections.

The special case of a quarter-wavelength spacing of insulators is of interest in that the characteristic impedance of the feeder is unmodified by the presence of the insulators. If the spacing of the insulators is in the neighbourhood of a half-wavelength the rate of change of the constants of the equivalent feeder is very rapid, and this condition is to be avoided. At this precise spacing the characteristic impedance goes to zero, which is associated with the whole capacitance of the insulators numerically adding.

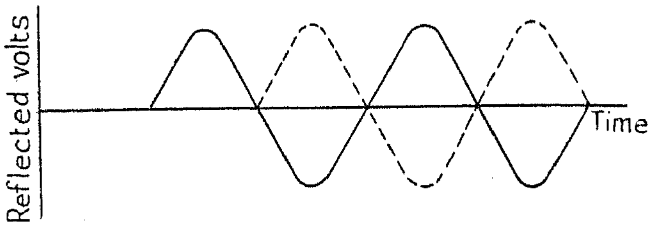

Fig. 14

An alternative method of considering the problem is in terms of a suddenly applied impressed sinusoidal voltage given by $e^{j\omega t}1$ in the notation of the operational calculus, applied to the feeder terminated by a condenser in parallel with its characteristic impedance. The reflected voltages arising are

$$-\frac{jZ_0C\omega e^{j\omega t}}{2 + jZ_0c\omega} - \frac{2e^{-\frac{2t}{Z_0c}}}{2 + j\omega Z_0c} \quad . \quad . \quad . \quad (16)$$

of which the second term is a transient of negligibly short duration and the first is a steady reflection. A second insulator placed farther down the feeder will give rise to a similar steady reflected wave, and if the insulators are spaced at $\frac{1}{4}$ wavelength the second wave will be delayed by a half-period and so these effects will cancel except for the first half-cycle. This is shown in Fig. 14, in which the full and dotted lines show the voltage waves reflected from the first and second insulators respectively. Hence if the insulators are spaced along the length of the feeder at $\frac{1}{4}$-wavelength intervals the resultant reflected wave will consist of a series of half-cycles as indicated in Fig. 15, which assumes an even number of insulators. The wave-form of Fig. 15 will persist for a time equal to the time of travel of the wave over twice the length of the feeder.

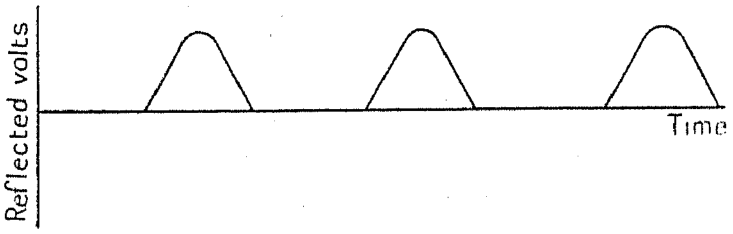

Fig. 15

The effect on the receiver is therefore to produce a step in the picture of this duration and of amplitude dependent on the amplitude of the carrier-frequency component of the wave of Fig. 15, in this case obviously about one-half that due to a single insulator. The relative amplitude of the wave reflected by an insulator of capacitance 1 $\mu\mu$F in an 80-Ω feeder at 45 Mc./sec. is $1 \cdot 1$ %, so that with the above spacing a step in the picture of $0 \cdot 5$ % amplitude would occur.

At closer spacings the cancellation of the carrier-frequency component is more complete. The spacing of half-wavelength is particularly objectionable in that the whole of the reflected waves add in phase.

If the spacing is not uniform there is a probability of considerably less perfect cancellation. For example, for completely random spacing of insulators the probable value of the amplitude of the reflected wave, corresponding to Fig. 15, would be $a\sqrt{N}$, where N is the number of insulators from which waves are arriving at any moment, and a the amplitude due to a single insulator.

The practical conclusion to be drawn from this discussion is that insulators of low capacitance should be used, spaced at small equal intervals.

(d) Bends and Expansion Joints

A difficulty in the construction of an ideal uniform feeder is that it is required to change its direction at certain points. At such bends considerable mechanical support of the inner is required to hold it in position, and it is usually necessary to provide additional insulators for this purpose. Such additional insulators give rise to irregularities of impedance. If the bend is in the form of an arc of a circle, a large number of insulators will be required and will produce a large impedance irregularity. This effect may be cancelled by such means as reducing the relative size of the inner at the bend so that with the additional capacitance of the insulators the characteristic impedance of the section remains unaltered. Alternatively, it is possible to cancel the effect of a group of insulators over a considerable range of frequency by a second similar group individually spaced $\frac{1}{4}$ wavelength away.

In the Alexandra Palace installation right-angle bends were adopted which permitted the use of angle boxes with a single large conical insulator for locating the inner at the bend. The boxes were so proportioned as to give a deficiency of capacitance at the box, which was partially compensated by the capacitance introduced by the insulator, and the residue was taken up by a simple variable plate condenser adjusted to the desired value.

In the construction of long lengths of feeder, expansion joints are provided in both the inner and outer conductors to allow for expansion with temperature-changes. These joints are frequently in the form of outer sleeves which are a sliding fit over the outer and inner of the feeder. Assuming an outer diameter of 5 in. and an inner of $1\frac{3}{8}$ in. with sleeves of $\frac{1}{32}$ in. wall, the change of the characteristic impedance is $4 \cdot 5\ \Omega$, which, for an expansion joint of length 5 ft. 6 in., gives the large amplitude of $\pm$ 9 ohms for the input-resistance variation. This fluctuation is due to the proportionally large variation of the inner diameter.

To reduce this cause of irregularity in the Alexandra Palace system, the sleeves over the inner were replaced by pipes sliding within the normal inner, which permitted the minimum length of abnormal-diameter section. The joints were located at the angle boxes so that the residual error of the joint could be compensated by the variable condenser located in the box.

(e) Eccentricity of the Feeder Conductors

The preceding Sections have described localized irregularities introduced into an otherwise uniform feeder. This uniform feeder itself is not realizable in practice on account of variations in the diameters, shapes, and degree of eccentricity, of the conductors. Of these factors eccentricity appears to be the most important in the Alexandra Palace type of feeder. The insulators were arranged to be sufficiently loosely fitting to permit the relative expansion of the inner conductor, and, since the conductors tend to be somewhat irregular in cross-section and the insulators must be sufficiently loose not to jam in the tightest parts, there is often a considerable amount of tolerance in the location of the inner conductor. As a result of this and the spacing between insulators, the inner conductor tends to be slightly bowed and eccentric in an irregular manner. This eccentricity produces impedance irregularities similar to those already discussed, but which, on account of their random nature, are much more difficult to reduce.

In the Alexandra Palace installation the other sources of irregularity were reduced to small values by the means already discussed, so that the eccentricity remained as the outstanding cause of irregularities in the feeder. It was found that, by rotating sections of the inner conductor, the input impedance could sometimes be changed by an ohm or so. To reduce the effect of eccentricity, the inner was rotated to a best position and then, if necessary, condensers were inserted at selected points along the feeder to produce counteracting irregularities in the manner previously discussed. Also the condensers in the angle boxes were used for this purpose.

Obviously it is very undesirable that this form of correction should be necessary, and, in an Appendix, an estimate is made of the tolerance which could be allowed in centring the inner conductor without excessive impedance irregularities arising. It is there concluded that, for a feeder similar to that at Alexandra Palace, if the displacement of the inner conductor from the centre could not exceed $0 \cdot 05$ in. the probable variation of input impedance due to eccentricity would be $\pm$ 1Ω and the greatest possible value could not exceed 5 ohms.

(f) Examples of the Variation of Input Impedance with Frequency of Matched Feeders

In this Section a few examples are given, illustrating the behaviour of actual feeders.

Fig. 16 shows an early measurement of the variation of the input resistance of a terminated feeder of length 152 ft. The feeder consisted of concentric copper pipes $3\frac{1}{4}$ in. and $\frac{7}{8}$ in. outer and inner diameters respectively, with porcelain bell-type insulators. It consisted of a straight portion of 116 ft. followed by three bends in the

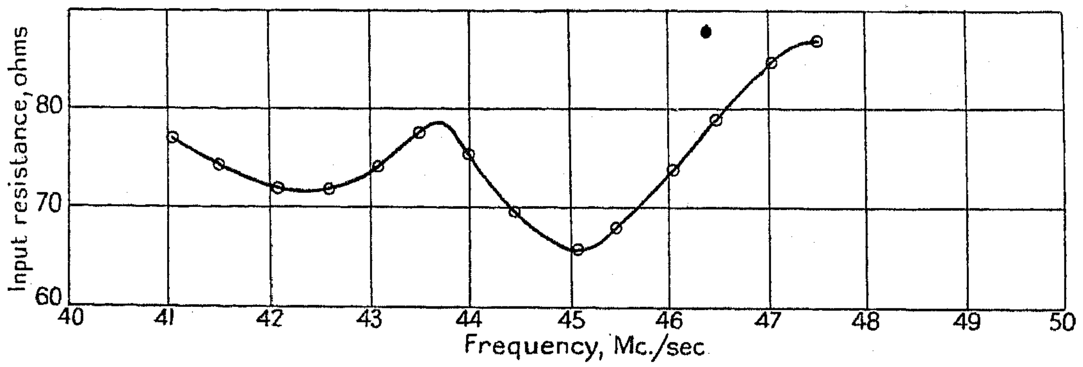

Fig. 16.—152-ft. length of 3¼ in. feeder with three right-angle bends.

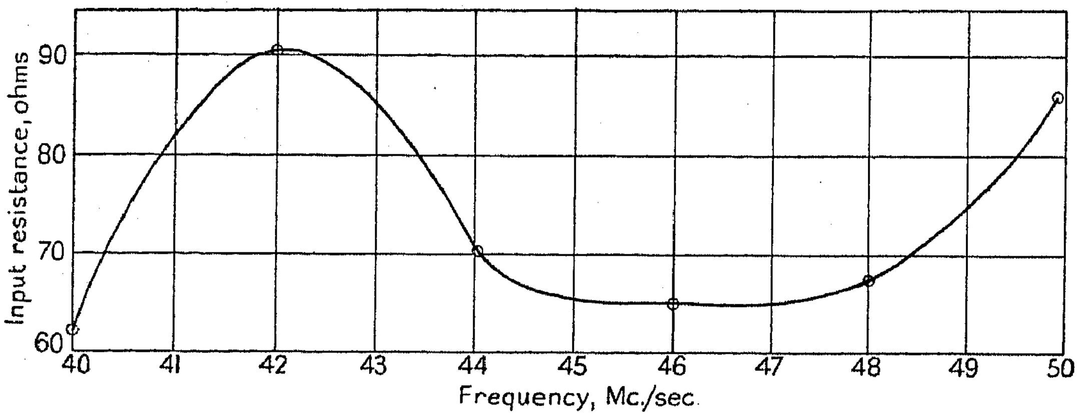

Fig. 17.—Straight 117-ft. length of 3¼ in. feeder with "bell" type insulators irregularly spaced.

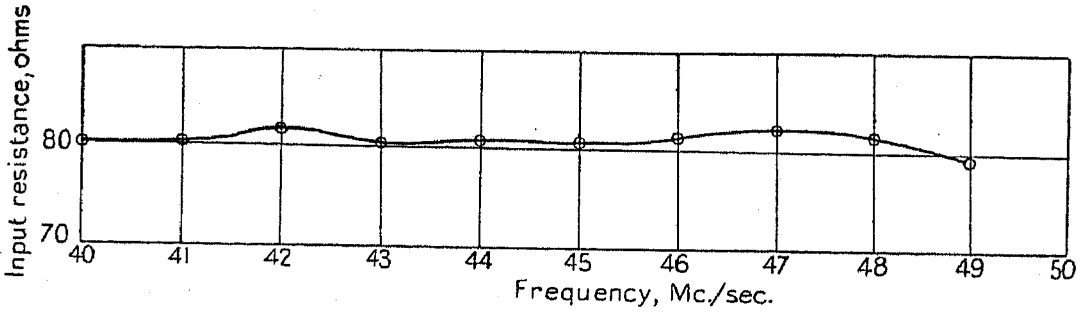

Fig. 18.—Straight 97-ft. length of 5-in. Alexandra Palace type feeder with steatite rod insulators at ⅛λ intervals.

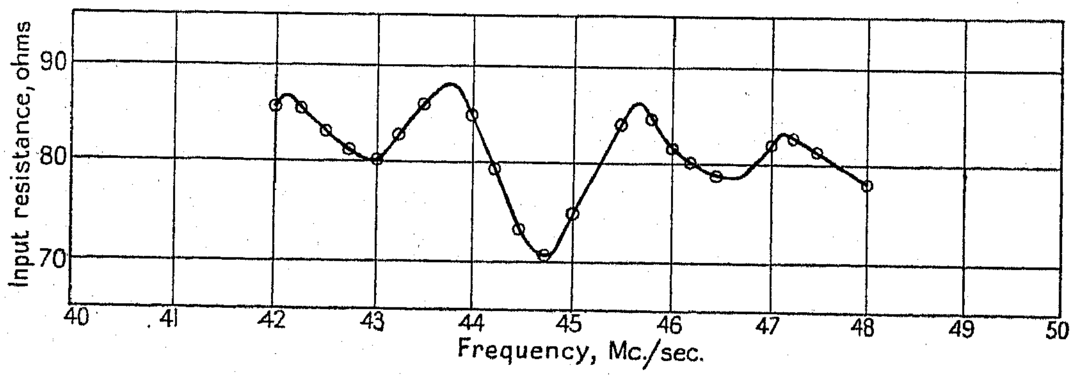

Fig. 19.—320 ft. of 5-in. Alexandra Palace type feeder without impedance correction.

form of arcs of circles and straight connecting portions. Insulators had been concentrated at each bend for mechanical reasons. The curve shows a fluctuation of resistance of considerable magnitude, suggesting a sine wave of increasing amplitude such as that given by two separated point disturbances.

Fig. 17 shows an input-resistance measurement on a similar straight section of feeder of length 117 ft. as originally laid for a sound transmitter in which no special precautions had been taken as to the spacing of the insulators.

Fig. 18 shows the results of the first 97-ft. trial section of feeder of similar type to that laid at Alexandra Palace. In this case low-capacitance insulators were used, spaced rigorously at equal intervals of about $\frac{1}{8}$ wavelength. The curve shows that the fluctuation is much reduced and of a periodicity that suggests a slight mistermination at the end of the feeder.

The Hayes vision feeder was laid with due attention to all the errors mentioned in the previous Sections, and the input impedance characteristic was measured as each additional section of the feeder was completed. Before proceeding to the next section of the feeder the variation of impedance was reduced to a small value by the adjustment of angle-box condensers, and in some cases by the insertion of additional condensers at appropriate places. As a result of this procedure the overall impedance characteristic varied less than $\frac{1}{2}$ ohm from the mean over the frequency range 41·5–47·5 Mc./sec. A feeder to the sound aerial was also in the course of construction and, as an experiment, it was decided to lay it as nearly as possible in the same manner as the vision feeder, except that no attempt was made to adjust the impedance characteristics of the individual sections in the course of construction. When completed an overall curve of impedance against frequency was taken. The result is shown in Fig. 19. The overall variation of resistance is large, being 17 ohms. The curve is irregular, and the obviously large component of "periodic interval" about 1·5 Mc./sec. indicates that there is a serious irregularity near the end.

It was concluded that it would be necessary for the vision system at Alexandra Palace to adjust individually each section of the feeder as had been done at Hayes.

(5) MATCHING THE AERIAL TO THE FEEDER

(a) The Aerial Transformer

The problem of matching the aerial, which is a resonant system whose impedance varies with frequency, to the feeder presents greater difficulty than those so far discussed. If the aerial is matched at the mid-frequency the deviations of impedance may cause a serious mismatch at the side-band frequencies.

In the Alexandra Palace system a transformer was provided to transform the aerial impedance to that of the feeder at the mid-frequency, and means were also provided to cancel partially these residual mismatches. Measurements of the aerial impedance, which was complex and about one-quarter of the characteristic impedance of the feeder, showed that both the resistive and reactive components varied with frequency.

Considerations of the possibility of simultaneously annulling the reactance and matching to the feeder at the mid-frequency led to the adoption of a transformer section of feeder, the principle of which is described below. Consider the arrangement in Fig. 20, where $Z_i = R + jX$ is the input impedance when the feeder is terminated by the impedance $Z_a = r + jx$. Then

$$Z_i = Z_0\left\{\frac{Z_a \cos\alpha + jZ_0 \sin\alpha}{Z_0 \cos\alpha + jZ_a \sin\alpha}\right\} \quad . \quad . \quad (17)$$

Separating equation (17) into its real and imaginary components and eliminating α, it can be shown that

$$R^2 + X^2 + Z_0^2 - R\frac{r^2 + x^2 + Z_0^2}{r} = 0 \quad . \quad (18)$$

Assuming, therefore, that we desire to transform to a given pure resistance R, we obtain, on putting $X = 0$,

$$Z_0^2 = Rr\left\{1 - \frac{x^2}{r(R - r)}\right\} \quad . \quad . \quad . \quad (19)$$

giving the required value for Z_0. We also obtain, by substitution in equation (17),

$$\tan 2\alpha = \frac{-\,2xZ_0}{Z_0^2 - r^2 - x^2} \quad . \quad . \quad . \quad (20)$$

We have therefore determined the constants of a feeder which, when connected in series with the aerial, transforms its impedance at one frequency to a given pure resistance. The operation of the feeder is somewhat similar to that of a $\frac{1}{4}$-wavelength transformer, transforming from a low resistance r to a high resistance R. Owing to the fact that it is terminated by a resistance with a small positive reactance, there is a length of feeder slightly shorter than a $\frac{1}{4}$ wavelength for which the input impedance is resistive. Further, because it is a considerably mismatched feeder it would, if terminated with a constant impedance, show a varying input reactance over the frequency range. This variation, in the case of the Alexandra Palace aerial, opposes that due to the change of the aerial reactance. In consequence the device tends to cancel the overall reactance of the aerial and to leave a substantially pure but varying resistance. For this reason, and on account of its simplicity, it was decided to make use of this type of transformer.

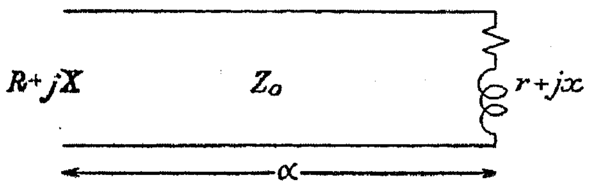

Fig. 20

(b) Reduction of the Residual Mismatch

Further to reduce the effect of the residual reflections at the side-band frequencies due to the variation of impedance with frequency after transformation, a method was adopted of deliberately introducing an irregularity at a chosen point in the feeder which would contribute a reflected wave in anti-phase to that due to the residual mismatch. The choice of the point of insertion and the nature of the irregularity to be inserted depends on the following considerations. To avoid loss of power it is desirable that the inserted irregularity should be purely reactive. Further, it is necessary that the inserted irregularity should have no effect at the frequency at which the aerial is matched to the feeder. In consequence a tuned circuit of low loss suggests itself as the possible form to be adopted. A parallel tuned circuit across the line would have a positive reactance at frequencies below the resonance, and vice versa. If, therefore, a point on the feeder exists at which the parallel resistance of the feeder terminated by the aerial-transformer section is constant, accompanied by a parallel reactance equal but opposite in sign to that of a tuned circuit, the connection of the tuned circuit across this point would cancel the reactance and leave a constant resistance. In this method deviations of the aerial impedance are transformed by the feeder to variations of reactance, which are then neutralized without loss of power by a tuned circuit of appropriate selectivity.

Consider a feeder, matched to the aerial at a frequency f_0, the characteristic admittance of which is represented by the vector OA of Fig. 21. Let the deviation of aerial admittance at a frequency f_1 be represented by the line AC, then a length α of feeder can be chosen for which the input conductance is equal to Y_0 and the susceptance is negative, being given by AP (see Fig. 21). This requires that $2\alpha = 2n\pi + \phi$, where n is any integer and ϕ is indicated in the Figure. If f_1 is greater than f_0 a tuned circuit resonant to f_0 can be found having a positive susceptance at f_1 of magnitude AP. In a similar manner, for a frequency f_2 equally below f_0 another series of lengths of feeder exist, having a conductance equal to Y_0 and a positive susceptance. This susceptance could be neutralized by the same tuned circuit if the deviation of admittance from Y_0 were equal to that at f_1. Since the electrical length of the feeder varies with frequency it is usually possible to find a physical length of feeder at the end of which a suitable tuned circuit would annul the susceptance at both frequencies.

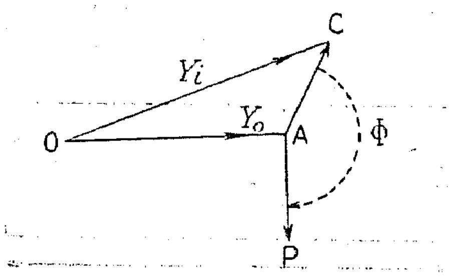

Fig. 21

As an illustration of the process, consider a case in which at both frequencies f_1 and f_2 the conductances are equal and lower than the characteristic conductance of the feeder, and the susceptances zero. In Fig. 22, OA represents the characteristic admittance of the feeder, and AC_1 and AC_2 are the deviations of the terminating conductance at frequencies f_1 and f_2 respectively. Assuming f_1 to be the higher frequency, it is required to find a physical length of feeder such that AC_1 rotates to AP_1 while AC_2 rotates to AP_2. It is seen that AC_1 must rotate through 180 degrees more than AC_2. This corresponds to the electrical length of the feeder being $\frac{1}{4}$ wavelength greater at the higher frequency. An alternative conception is that if waves of frequencies f_1 and f_2 are applied to the required length of feeder the waves reflected from the end at frequencies f_1 and f_2 must arrive in anti-phase. In this particular case, since the reflecting impedances are identical the electrical length of the feeder at the two frequencies must differ by $\frac{1}{4}$ wavelength. This length is given by

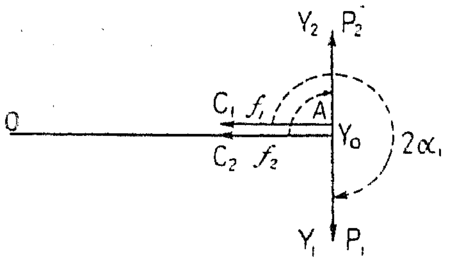

Fig. 22

$$l = \frac{\lambda_0}{4}\cdot\frac{f_0}{f_1 - f_2}$$

If $f_0 = 45$ Mc./sec. and $(f_1 - f_2) = 4$ Mc./sec., $l = \frac{1}{4}(11{\cdot}25\lambda)$. For this length P_1P_2 is not at right angles to OA but the nearest value to obtain this relationship is 11 quarter-wavelengths, and for this length AP_1 and AP_2 are still substantially in anti-phase.

Let a parallel tuned circuit be connected across the feeder so as to tune at f_0 and of such selectivity as to have susceptances $-AP_1$ and $-AP_2$ at frequencies f_1 and f_2 respectively. A convenient form for such a circuit consists of a $\frac{1}{4}$-wavelength of feeder short-circuited at one end and tapped at an intermediate point AB as shown in Fig. 23. For this circuit the rate of change of admittance with frequency from the tuned value is given by

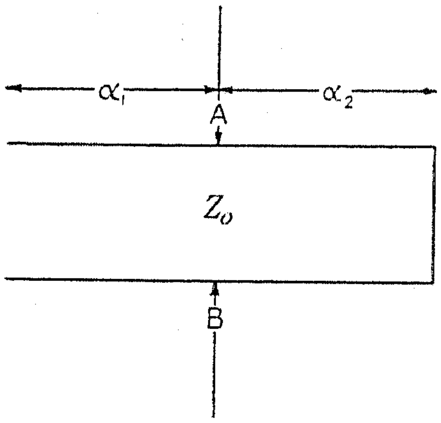

Fig. 23

$$\frac{dY}{df} = j\,\frac{\pi \sec^2 \alpha_1}{2fZ_0} \quad . \quad . \quad . \quad . \quad (21)$$

so that for a given selectivity and Z_0 the length α_1 may be determined.

From the above it is evidently possible to find a tuned circuit which when placed across the feeder at a selected point will reduce to zero, at two frequencies as well as at the matched frequency, the variation due to the varying aerial impedance. For other frequencies in the range the cancellation may not be perfect, but for simple types of variations of the aerial impedance it will be found possible to reduce substantially the overall fluctuations by this means.

(6) IMPEDANCE-MEASURING TECHNIQUE

(a) Measuring Apparatus

Special forms of impedance-measuring gear were devised, since at these frequencies considerable difficulty is experienced with the appreciable reactance of even the shortest leads, the large effect of small stray capacitances, and the lack of reliable standard resistances. One form was self-contained and portable and capable of measuring resistance and reactance over a wide range of values of impedance and frequency. A second form was designed to attach to the end of the feeder, and was capable of measuring small deviations of resistance from a fixed value.

In both forms the impedance to be measured is placed in parallel across a tuned circuit which is coupled to an oscillator and has a diode voltmeter across the condenser. The circuit is tuned by means of the condenser, and the diode volts are noted. A resistance is then substituted in place of the unknown impedance of such a value as to give the same diode volts when the circuit is retuned. The resistance is equal to the parallel resistance of the impedance, and the change in capacitance of the resonant circuit measures the parallel reactance. The apparatus is enclosed in a copper case to which one side of the tuned circuit is connected.

The portable gear (see Fig. 24*a*) comprised a variable-frequency oscillator followed by an amplifying and isolating stage, the anode of which was tuned. To this tuned circuit the measuring circuit was inductively coupled, provision being made to switch coils in parallel with the measuring circuit in order to change the point of tuning on the condenser. An Isolantite low-loss 150-$\mu\mu$F condenser calibrated at low frequencies was used as the measuring condenser. A method of measuring the residual inductance was devised, so that the equivalent values of the capacitance at high frequencies were known. The method consisted in noting the apparent value of a small fixed condenser placed in the measuring cups at various points over the range of the variable condenser. The condenser setting for resonance was varied by placing other fixed reactances in parallel across the tuned circuit. From the variation of these apparent values the effective inductance of the condenser was calculated and the equivalent capacitances at high frequencies were deduced. This method has since been described by Field and Sinclair.* Across the tuned circuit a diode voltmeter of variable sensitivity was connected. Mercury cups were provided as terminals, and leads to the case and the terminals were made as thick and short as possible. The apparatus was mains-operated.

The feeder-measuring gear (see Fig. 24*b*) was adapted for measuring accurately small changes in the feeder input resistance. The layout of the measuring tuned circuit was modified to reduce the length of feeder connecting leads to a minimum. The inductance of the tuned measuring circuit was formed of a pick-up coil and a suitable length of screened connecting cable so that the oscillator could be removed to a distance of a few feet. A box containing the measuring condenser, the diode, and its associated circuits, was clamped on to the end of the feeder and so arranged that a short sleeve over the inner of the feeder could be slipped down into a mercury pool. A switch of the rocking type was provided, connected to the live terminal of the condenser and dipping either into the mercury pool or into a second pool, provided to permit the bridging of a suitable resistance to the case. This switch, on account of its finite inductance, introduced a small error into the measured impedance so that this gear could only be used for the comparison of substantially equal resistances. The diode voltmeter was arranged as before but with the addition of a circuit for backing off the steady d.c. diode current and the provision of a suitable galvanometer to obtain enhanced sensitivity.

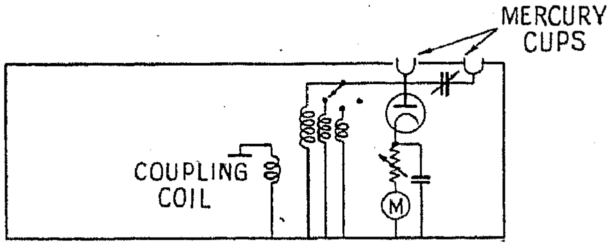

Fig. 24(*a*).—Portable impedance gear.

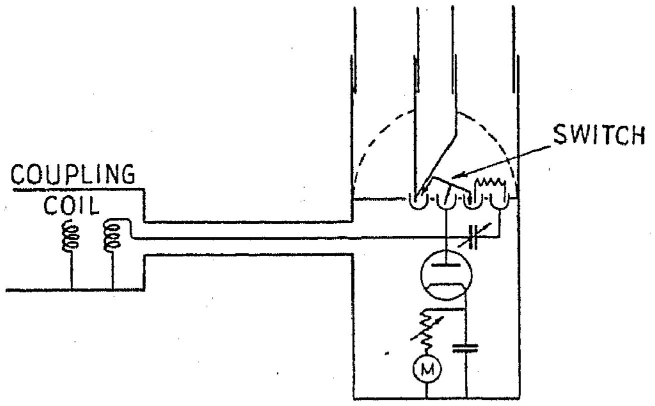

Fig. 24(*b*).—Feeder-measuring gear.

The accuracy of the frequency calibration is of considerable importance as the measured variation of resistance with frequency has a period of the order of 1 Mc./sec. If, therefore, we wish to insert an irregularity to compensate for an observed error, it is necessary to know the frequency to an accuracy of about 0·05 Mc./sec., in order accurately to locate the point of insertion from the phase of the oscillation of the input resistance. The oscillator was arranged to have a very open scale, and the calibration was obtained from the transmitter frequency (45 Mc./sec.) and by short-circuiting the feeder at a distant point and measuring the $\frac{1}{4}$-wavelength frequencies which gave known frequency-differences.

Fixed resistors of the Dubilier half-watt metallized type were used, and intermediate values were obtained by interpolation and by the use of parallel combinations. The d.c. values were measured after use and were accepted as correct at high frequencies. This procedure was justified by the fact that the characteristic impedance of the terminated feeder came to the theoretical value when measured in this way.

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* *Proceedings of the Institute of Radio Engineers*, 1936 vol., 24, p. 255.

Using the portable gear, measurements of resistance of the order of 80 Ω could be reproduced to about 0·5 Ω with an absolute accuracy of about 1 or 2 Ω, and measurements of capacitance to about 1 μμF. This capacitance corresponds to a parallel reactance of 3 600 Ω or a series reactance of 1·7 Ω. The feeder-measuring gear was adjusted to give small differences in resistance correct to 0·1 Ω, and the capacitance change was not recorded.

(b) Feeder-Terminating Resistors

Terminating resistors were formed of short sections of line (as shown in Fig. 25) which plugged on to the end of the feeder. A number of half-watt resistors arranged in parallel were connected by short leads across the end of the feeder, the residual inductance of the leads being cancelled by a small variable condenser C. The inner of the feeder was located by means of insulator rods passing through the inner and outer. It was necessary to adjust the resistor and condenser, so that in the presence of the rod insulators the input impedance across B was equal to the characteristic impedance of the main feeder. This adjustment was performed by preceding the terminator by various sections of feeder such as $\frac{1}{8}$, $\frac{1}{4}$, $\frac{3}{8}$, $\frac{1}{2}$ wavelength, etc., in succession, and adjusting values until equal input resistances were measured on each section. The choice of these particular lengths of feeder depends on the fact that the input resistance is a function of feeder length if misterminated.· The departure from the characteristic impedance for the $\frac{1}{8}$ and $\frac{3}{8}$ wavelengths are equal and of opposite sign and depend mainly on the reactance error in the termination. It is therefore possible to adjust the reactance by resetting the condenser to give an input resistance which is the mean of the previous values. In a similar manner the $\frac{1}{4}$ and $\frac{1}{2}$ wavelengths may be used to determine the adjustment of the terminating resistance. The terminator could be adjusted in this manner at 45 Mc./sec. to an accuracy of 0·2 %.

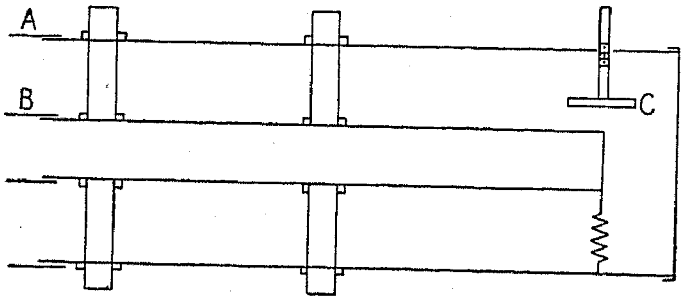

Fig. 25

A series of measurements taken at 43, 45, and 47 Mc./sec. for the four lengths, in order to test the selectivity of the arrangement, are shown in Table 1. The figures indicate that the terminating resistance was 0·4 Ω too low but was non-reactive, since the $\frac{1}{2}$-wavelength results are consistently low, and the $\frac{1}{4}$-wavelength results high by this amount. There appears to be an uncertainty of measurement of about 0·2 Ω over this frequency range, but no systematic variation.

The resistors are unfortunately subject to change with temperature and to occasional spasmodic alterations. In the original work at Hayes it was found necessary to provide means to keep the terminator at an approximately constant temperature. Later, low-temperature-coefficient resistors were obtained and the heating control was dispensed with. A constant check was maintained on the terminator during the measurement by measuring the d.c. resistance.

Table 1

INPUT RESISTANCES R_i IN OHMS TO LENGTHS OF FEEDER SPECIFIED WHEN TERMINATED BY "TERMINATING RESISTANCE"

Frequency	43 Mc./sec.		45 Mc./sec.		47 Mc./sec.	
Length of feeder	R_i	$R_i - Z_0$	R_i	$R_i - Z_0$	R_i	$R_i - Z_0$
$\frac{1}{8}\lambda$	77·6	0	77·8	0·2	77·6	0
$\frac{1}{4}\lambda$	78	0·4	78 0	0·4	78·15	−0·55
$\frac{3}{8}\lambda$	77·6	0	77·6	0	77·9	0·3
$\frac{1}{2}\lambda$	77·3	−0·3	77·1	−0·5	77·3	−0·3

(7) THE VISION SYSTEM AT ALEXANDRA PALACE

(a) General Arrangement

The radiating system finally adopted consisted of two rings of full-wave dipoles with the mast in the centre as illustrated diagrammatically in Fig. 26, in which Fig. 26(*a*) is a plan view and Fig. 26(*b*) a false section along the line APE. It is seen that the main feeder is brought to a central point P, from which radiate lines to the various aerial units which are indicated by the

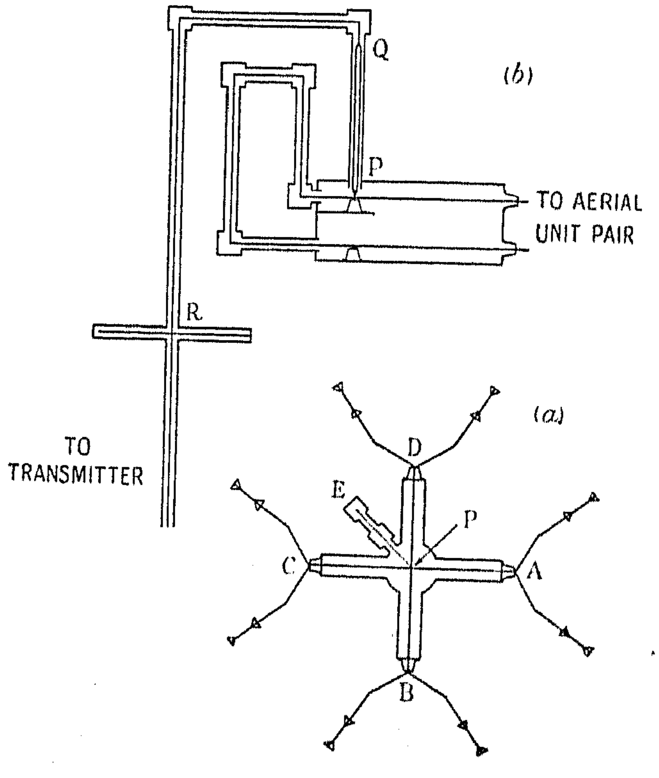

Fig. 26.—Arrangement of aerial and feeder connections.

triangles of Fig. 26(*a*). The change from the unbalanced feeder to the balanced aerial system is accomplished by means of a half-wave phase-reversing loop. From the point P (Fig. 26), the point of connection of the feeder to the aerial, a transforming section of approximately $\frac{1}{4}$ wavelength is used to transform the complex aerial impedance to the characteristic impedance of the feeder at the carrier frequency. Beyond this point the main feeder with normal inner conductor runs 450 ft. to the transmitter. A correcting circuit is inserted in parallel with the feeder at approximately 60 ft. from P to reduce the mismatch at the side-band frequencies, and at the transmitter a $\frac{1}{8}$-wavelength transformer is used to transform the characteristic impedance to $50\,\Omega$, which is more suitable for loading the transmitter.

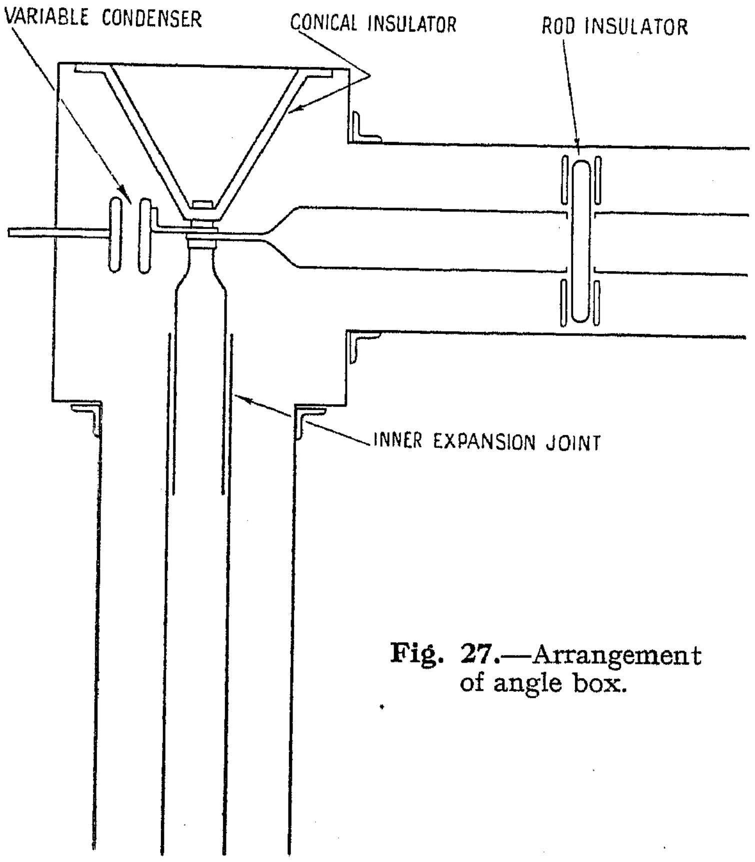

Fig. 27.—Arrangement of angle box.

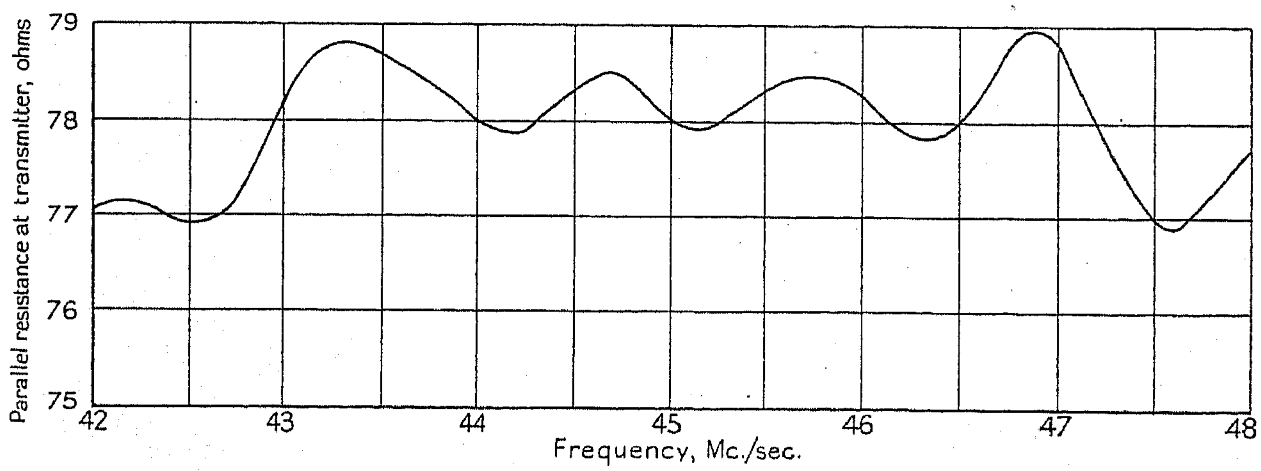

Fig. 28.—Impedance/frequency characteristic of the Alexandra Palace vision feeder, terminated with Z_0 4 ft. below R (Fig. 26).

(b) The Feeder

The main vision feeder runs from the vision aerial platform vertically down to the base of the steel mast, to which it is bonded throughout its length. It passes to the outside of the supporting tower, vertically down to the colonnade, and then about 100 ft. horizontally to a change-over box in the wall of the sound transmitter room. From this box two equal branches lead to the two vision transmitters originally installed in Alexandra Palace. The sound feeder runs beside the vision feeder. Both feeders are laid with as few bends as possible, and those which do occur are right-angle bends in the form of angle boxes.

The feeders consist of concentric copper pipes of 5 in. and $1\frac{5}{8}$ in. diameter, respectively. The characteristic impedance of such a feeder if air-spaced would be $78\,\Omega$ and the attenuation 1 neper in 4·8 miles. The inner conductor is located by means of steatite low-capacitance insulators in the form of rods passing through the inner and having sleeves slipped over the ends of the rods of such a length as to centralize the inner (see Fig. 27). The rods are $\frac{3}{8}$ in. diameter and the sleeves $1\frac{3}{4}$ in. long. The insulators were spaced at equal intervals of $\frac{1}{8}$ wavelength at 45 Mc./sec., alternately at right angles, and ferrules were sweated into the holes in the inner to act as guides for the rods and to increase the bearing surfaces. The capacitance introduced by one insulator was $0\cdot4\,\mu\mu\mathrm{F}$, and the reduction of characteristic impedance given by equation (14) was $0\cdot35\,\Omega$. The loss was extremely small and was not measured.

A source of trouble with feeders of this type is the condensation of water on the insulators, which can cause losses and introduce impedance irregularities in the feeder. This was encountered in one of the feeders laid at Hayes, but was absent in subsequently laid feeders. Provision was made at Alexandra Palace for drying the feeder, should it prove necessary, by passing a large 50-cycle current through it.

The type of angle box used is shown in Fig. 27. It consists of an approximately cubic brass box containing a large conical insulator and a variable condenser in the form of a disc on the end of a threaded spindle. The same Figure shows an inner expansion joint located at the box.

The feeder was laid from the transmitter end, and at the conclusion of suitable sections, usually at an angle box, the feeder was terminated and the variation of input resistance with frequency measured. Adjustments were made, as in the case of the Hayes vision eeder, to reduce the variation. The final curve of resistance against frequency for the 390 ft. up to the point of insertion of the tuned circuit is shown in Fig. 28. It will be seen that the variation between 43 and 47 Mc./sec. was not greater than ± 0·5 Ω.

At the transmitter end the arrangement transforming the characteristic impedance to 50 Ω consisted of a condenser formed by a short length of open-circuited feeder branching from the main feeder about $\frac{1}{8}$ wavelength from the transmitter.

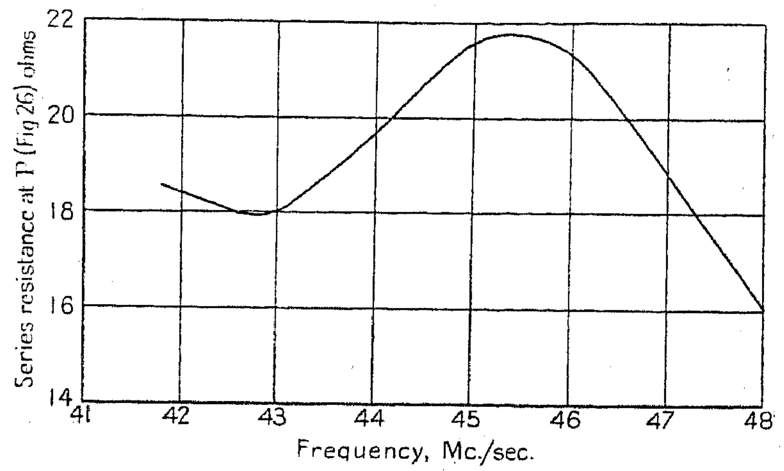

Fig. 29(*a*).—Resistance at P (Fig. 26).

(c) The Aerial Transformer

The design of this transformer is based on the measurements of the impedance between the point P (Fig. 26) and the case. For this purpose a $\frac{1}{4}$-wavelength at 45 Mc./sec. of normal feeder was connected to the point P and the impedance over a range of frequencies was measured by means of the portable impedance gear. The impedances at P at the various frequencies were deduced from these measurements. Figs. 29(*a*) and 29(*b*) show the series resistance and reactance at the point P, plotted against frequency. The curves show a considerable reactive term and a variation of both resistance and reactance with frequency. These variations are not of simple form and are due to the combined effects of the selectivity of the aerial elements themselves and the various small mismatches occurring in the distribution system.

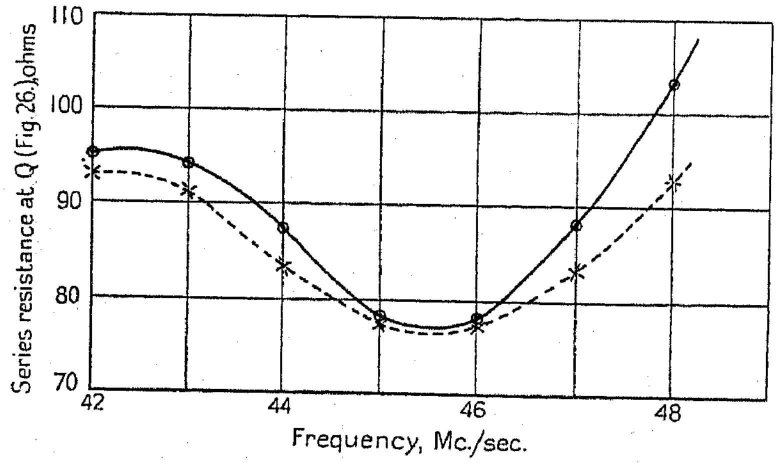

Fig. 30(*a*).—Resistance at Q (Fig. 26).

— o — Values obtained after insertion of transformer section PQ.
- - X - - Values predicted from initial measurements.

In accordance with the principles discussed in Section 5(*a*), a transforming length of feeder with normal outer but $2\frac{9}{16}$ in. diameter inner conductor and 4 ft. 6 in. long was inserted to transform the impedance at 45 Mc./sec. to the 78 Ω of the feeder. The impedance after inserting this transformer was measured in a similar manner to that at point P. Figs. 30(*a*) and 30(*b*) show the resulting values of resistance and reactance over the same frequency range, together with those predicted from the initial measurements and the constants of the transformer. It is seen that there is substantial agreement between the two sets of values, indicating that the measurements were reliable. The impedance is resistive and matched to the feeder at 45 Mc./sec., but there is a substantial variation at the side-band frequencies.

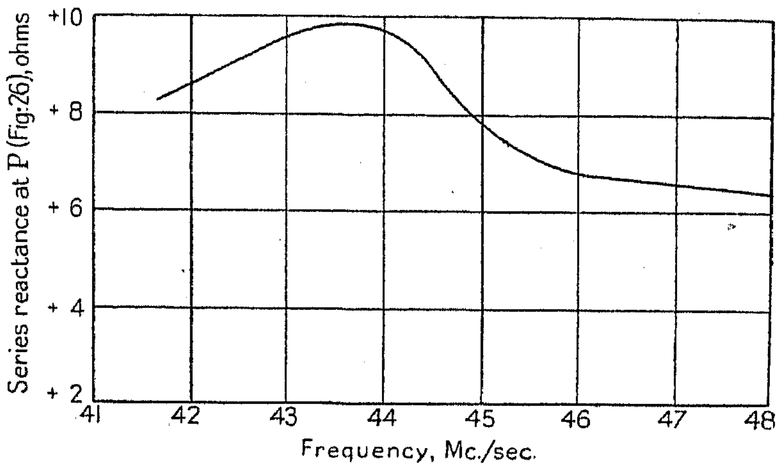

Fig. 29(*b*).—Reactance at P (Fig. 26).

(d) The Correcting Section and Overall Curve

In accordance with the principles discussed in Section 5(*b*), an attempt was made to reduce the effects of this variation by the insertion of a tuned circuit at an appropriate point in the main feeder. From the values given in Fig. 30 it was decided that the best point for the insertion of the circuit was $2\frac{1}{4}$ wavelengths at 45 Mc./sec. beyond point Q. The calculated values of parallel resistance and capacitance at this point are shown in Figs. 31(*a*) and 31(*b*). The resistance is seen to be roughly constant over most of the frequency range, and the capacitance variation is such that it could be

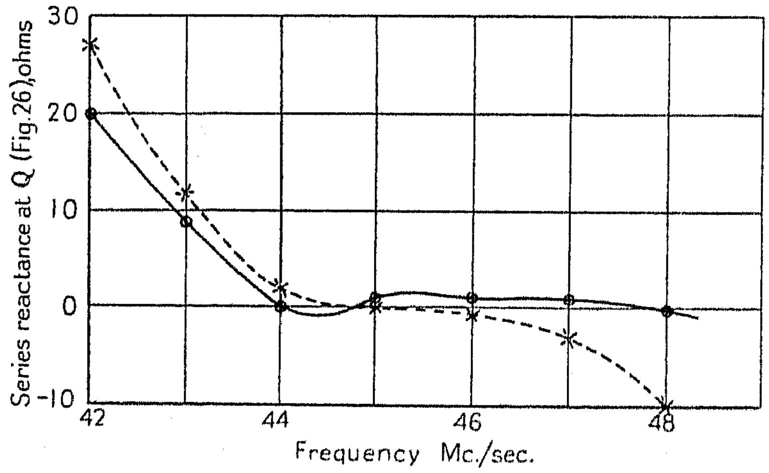

Fig. 30(*b*).—Reactance at Q (Fig. 26).

— o — Values obtained after insertion of transformer section PQ.
- - X - - Values predicted from initial measurements.

substantially reduced by the insertion of a tuned circuit. The feeder was cut in the vicinity of this point, and the impedance was measured at several nearby points. The measured results at the finally selected point of insertion, which was as nearly as could be measured the predicted point, are also shown in Figs. 31(a) and 31(b). In this case the resistance values agree reasonably, but there is a consistent error of a few micro-microfarads in the reactance measurements.

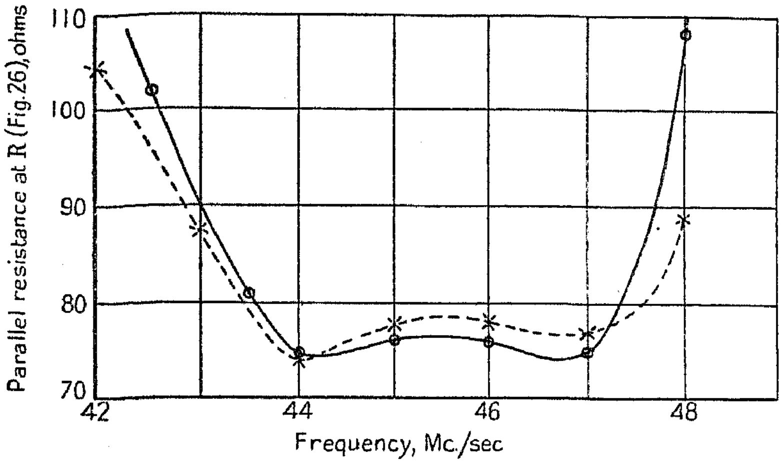

Fig. 31(a).—Resistance at R (Fig. 26).

— o — Values measured at point R.
- - X - - Values calculated from measurements of Fig. 30.

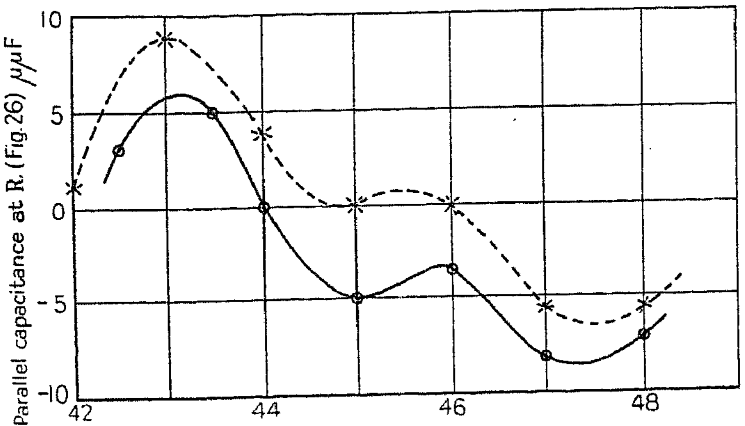

Fig. 31(b).—Capacitance at R (Fig. 26).

— o — Values measured at point R.
- - X - - Values calculated from measurements of Fig. 30.

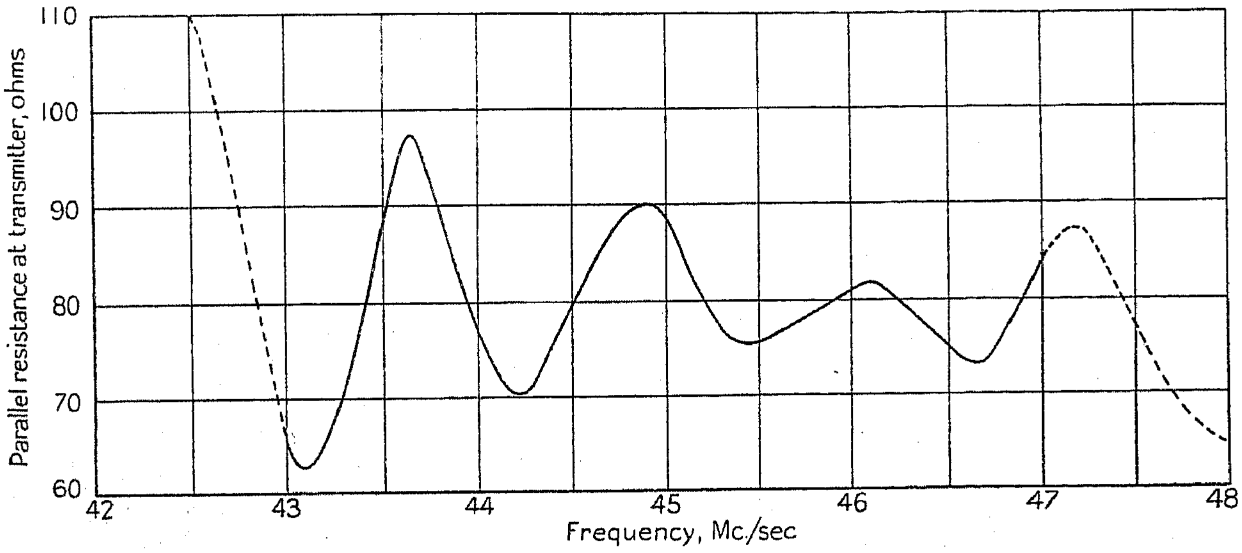

Fig. 32.—Impedance/frequency characteristic of aerial feeder system without parallel $\frac{1}{4}\lambda$ connection.

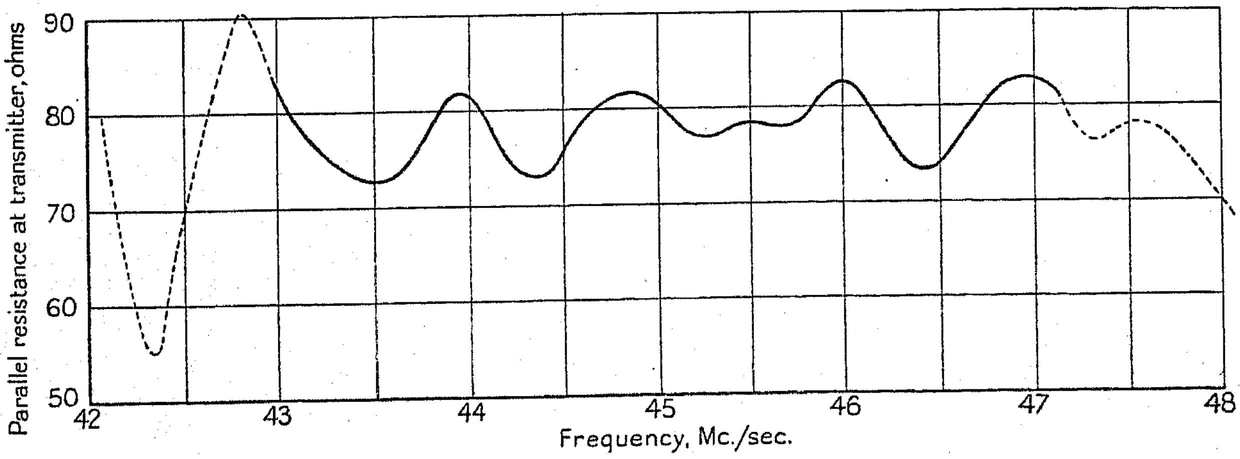

Fig. 33.—Impedance/frequency characteristic of Alexandra Palace vision aerial-feeder system.

Both curves agree, however, in indicating that the tuned circuit to annul the capacitance change should be of such a selectivity to give a capacitance change of 3 $\mu\mu$F per megacycle. The tuned circuit consists of a transverse section of normal feeder $\frac{1}{4}$ wavelength long, one end being open and the other short-circuited. Both ends are capable of sliding extension. With this arrangement both the tuning and the selectivity can be varied in situ. For the selectivity required, equation (21) gives for the length of the open end $\alpha_1 = 43°$, so the open and closed arms were made approximately equal. This form of tuned circuit was suggested by Mr. Rust, of the Marconi Co.

At this stage of the construction the vision feeder was completed from the transmitter to this point, and it was known that it was substantially free from impedance irregularities. It was decided to make the subsequent measurements from the transmitter end of the feeder, at which point it would only be necessary to measure resistance variation with frequency. Fig. 32 shows the measured variation of input resistance with frequency, the feeder being connected through without the tuned circuit in position.

Fig. 33 shows the same quantity after the tuned circuit had been inserted, adjusted to its optimum setting, and a slight adjustment made to the condenser in one of the upper angle-boxes. It is seen that the overall variation over the frequency range 43 to 47 Mc./sec. is not greater than ± 5 Ω.

It is possible that this variation might have been further reduced, but considerable difficulty was experienced in making the measurements up the mast under the unfavourable weather conditions prevailing. Further, time was limited by the requirement that the aerial should be complete for preliminary tests on the station. Moreover, the investigations at Hayes indicated that this amount of variation of impedance would lead to no visible distortion.

On completion of the station, observations on the transmitted picture showed no trace of striae due to multiple reflections in the feeder.

(8) ACKNOWLEDGMENTS

These experiments have been carried out in the Research Laboratories of Electric and Musical Industries, Ltd., Hayes, and the authors wish to express their indebtedness to Mr. I. Shoenberg, Director of Research, and to Mr. G. E. Condliffe and the staff of the Laboratories for their encouragement and assistance in the investigation.

In particular, thanks are due to Mr. A. D. Blumlein, who foresaw the effects of a mismatched feeder on the transmitted picture.

This paper deals with the feeder portion of the radiating system at the London Television Station. The aerial portion was designed and installed by the engineers of Marconi's Wireless Telegraph Co., and acknowledgments are due to Mr. Tringham and Mr. O'Neill for their collaboration throughout the whole of the work.

APPENDIX

The Input Impedance of an Approximately Terminated but Irregularly Eccentric Feeder

(a) Formulae for the input impedance as a function of eccentricity.

In this Appendix expressions are derived for the input impedance of a terminated, imperfectly "concentric" feeder in terms of the displacement of the inner conductor from the central position. This displacement is supposed to vary throughout the length of the feeder.

Consider the feeder (see Fig. 34) to be split up into a number of sections each sufficiently short for the eccentricity, and hence the characteristic impedance, to be assumed constant in the section. By an extension of the method of Section 3(*b*), which gave for a single section of the feeder of abnormal characteristic impedance Z_1 the expression

$$Z_i = Z_0 + 2j \sin \gamma_1 (Z_1 - Z_0) e^{-2j\alpha_1} \quad . \quad . \quad (22)$$

it can be shown that for the complete feeder of Fig. 1 the input impedance Z_i is given by:—

$$Z_i = Z_0 + \Sigma 2j \sin \gamma_1 (Z_1 - Z_0) e^{-2j\alpha_1} \quad . \quad . \quad (23)$$

where α_1 and γ_1 are the electrical lengths indicated in the Figure, and the summation includes each section into which the feeder is divided.

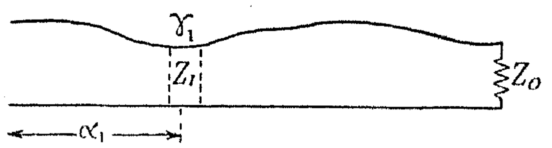

Fig. 34

By dividing the feeder into n equal sections of length γ this formula gives the Fourier series

$$Z_i = Z_0 + \sum_{m=1}^{m=n} 2j \sin \gamma (Z_m - Z_0) e^{-2j(m-\frac{1}{2})\gamma} \quad . \quad (24)$$

or by making the sections infinitesimal so that $\gamma = d\alpha$ the formula gives the Fourier integral

$$Z_i = Z_0 + \int_0^{\alpha_1} 2j(Z - Z_0) e^{-2j\alpha} d\alpha \quad . \quad . \quad (25)$$

where Z is the characteristic impedance at a length α from the input end and α_1 is the total length of the feeder.

The next consideration is to determine the values of the characteristic impedance in terms of the eccentricity of the conductors. Consider two cylinders of radii r_0 and r_1, one surrounding the other and having their axes separated by a distance d, as shown in Fig. 35.

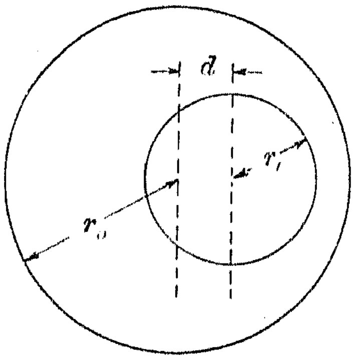

Fig. 35

For this arrangement with air spacing, Moullin* gives

$$C = \frac{0 \cdot 24}{\log_{10} \beta + \sqrt{(\beta^2 - 1)}} \, \mu\mu\text{F/cm.} \quad . \quad . \quad (26)$$

where

$$\beta = \frac{r_0^2 + r_1^2 - d^2}{2 r_0 r_1} \quad . \quad . \quad . \quad . \quad (27)$$

From this it can be shown that the change in Z_0 due to a small displacement d from the centre is given by

$$\Delta Z_0 = - \frac{60 d^2}{r_0^2 - r_1^2} \quad . \quad . \quad . \quad . \quad (28)$$

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* "Radio-Frequency Measurements," 1931, p. 291.

and for a feeder for which

$$2r_0 = 5 \text{ in.}, \quad 2r_1 = 1\tfrac{3}{8} \text{ in.}$$

$$\Delta Z = -\,10{\cdot}4d^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (29)$$

where d is in inches and ΔZ in ohms.

(b) Permissible limits of eccentricity.

It is desirable to estimate the tolerances which may be allowed in centring the inner conductor without the introduction of input-impedance variations greater than a given permissible value. The problem does not admit of an exact solution, since if the only information available is that the mechanical constraint prevents a displacement greater than a given value, the actual variation of the displacement, and with it the input impedance, may assume any number of forms. In order to arrive at a useful conclusion two possible configurations of the inner conductor, subject to a given maximum displacement, are assumed.

In the first case the configuration is that giving the greatest possible deviation of the input resistance from the characteristic impedance of the feeder (Z_0) when the conductors are concentric. From the equation

$$r_i = Z_0 + 2\int_0^{\alpha_1}(Z - Z_0)\sin 2\alpha d\alpha \quad . \quad . \quad (30)$$

and $$Z - Z_0 = -\,10{\cdot}4d^2$$

which applies to this particular type of feeder, it is seen that if d is constrained to be less than d_0 then $(Z - Z_0)$ may vary between 0 and $-\,10{\cdot}4d_0^2$. The integrand of equation (30) is the product of two variables $(Z - Z_0)$ and $\sin 2\alpha$, the former being a function of the configuration. The greatest value of the definite integral will be given by the configuration which gives a maximum value of $(Z - Z_0)$ when $\sin 2\alpha$ is positive, and a minimum value when $\sin 2\alpha$ is negative. In this case, with alternate $\frac{1}{4}$-wavelengths of the two extreme characteristic impedances Z_0 and $(Z_0 - 10{\cdot}4d_0^2)$, the value of $(r_i - Z_0)$ is $2 \times 10{\cdot}4d_0^2 2x/\lambda$, where x is the length of the feeder. This gives for a permissible resistance deviation of 1 ohm on a feeder of 20 wavelengths the value $d_0 = 0{\cdot}035$ in.

The above peculiar configuration of the inner is obviously unlikely to occur in practice, and the tolerance obtained is unnecessarily stringent. A second type of configuration is therefore treated in which the eccentricity within the given limits is assumed to be of a random nature, and a probable value of $(r_i - Z_0)$ is obtained. Serious difficulties arise in specifying the probability of any given configuration, because, owing to the rigidity of the pipe forming the inner, the eccentricity of adjacent sections of the feeder is not independent. In consequence the following treatment is not rigorous, but it is hoped that it would give the right order of magnitude. The feeder is supposed to be divided into n equal sections of length l. This length is chosen such that the eccentricity is approximately constant over one section, but the neighbouring sections are substantially independent. From observations of the degree of bending of the inner pipe these sections were estimated to be of length 2 ft. The final value of $(r_i - Z_0)$ obtained from equation (24) is given by

$$r_i - Z_0 = \sum_{m=1}^{m=n} 2(Z_m - Z_0)\sin\frac{4\pi l}{\lambda}(m - \tfrac{1}{2})\sin\frac{2\pi l}{\lambda} \quad . \quad (31)$$

where Z_m is the characteristic impedance of the mth section. The most probable value P for the sum of this series is the square root of the sum of the squares.

Hence

$$P = \sqrt{\left[\sum_{m=1}^{m=n}\left\{2(Z_m - Z_0)\sin\frac{2\pi l}{\lambda}\right\}^2\right]} \quad . \quad (32)$$

$$= \sqrt{n}\, 2a\sin\frac{2\pi l}{\lambda} \simeq \sqrt{n}\frac{4\pi la}{\lambda} \quad . \quad . \quad . \quad . \quad (33)$$

where a is the r.m.s. value of $(Z - Z_0)$. If x is the total length of the feeder $n = x/l$ and

$$P = 4\pi\frac{\sqrt{(xl)}}{\lambda}a \quad . \quad . \quad . \quad . \quad (34)$$

The total variation with frequency of r_i is of the order of $\pm P$. Equation (34) indicates that P increases with the feeder length x, as would be expected. The increase of P with the section length l corresponds to the fact that a small number of terms of random phase tend to give a sum greater than a large number of smaller terms which average to more nearly zero. It follows that the non-uniformity of a feeder as measured by P may be reduced by using a more flexible inner conductor, so reducing l, provided that it can be made central within the same maximum limits.

Equation (34) is capable of experimental verification. The steatite rod insulators allow a displacement d parallel to their length of $\frac{1}{32}$ in. and perpendicular to this of $\frac{1}{4}$ in., and the pipe tends to bend appreciably in the 5 ft. 6 in. interval between similarly orientated insulators. Consequently the average displacement may be taken as roughly $0{\cdot}15$ in. This gives a value of a of $10{\cdot}4 \times 0{\cdot}15^2 = 0{\cdot}23\ \Omega$. Taking l as 2 ft., then at 45 Mc./sec. equation (34) becomes

$$P = 0{\cdot}18\sqrt{x}\ (x \text{ in feet}) \quad . \quad . \quad . \quad (35)$$

Table 2 gives the mean variation of the input resistance from Z_0 for a number of lengths of this type of feeder which were measured at Hayes, and for the variation of which no other cause than eccentricity is known. Values of P calculated from equation (35) are also included. Since the agreement between the theoretical and measured results is fair, being no worse than the

Table 2

Length of feeder, ft.	79	320	216	85	70	80
Mean deviation from Z_0, Ω . .	0·7	5	2	1	2	0·5
P (calculated), Ω	1·6	3·2	2·6	1·7	1·5	1·6

random variations in the measurements, there is some experimental justification for the theory.

For a permissible value of P of $1\,\Omega$ and $x = 450$ ft. and $l = 2$ ft., equation (34) gives $a = 0\cdot06\,\Omega$; the corresponding displacement d from the central position is $0\cdot07$ in. For this maximum displacement the greatest possible value of the deviation of the input impedance from Z_0 due to the eccentricity is $5\,\Omega$. It would appear, therefore, that in the design of future feeders of this nature the trouble due to eccentricity would not be serious if the displacement of the inner conductor could be restrained from exceeding about $0\cdot05$ in. For greater lengths of feeder the tolerance would be correspondingly restricted.

Recently two papers* have appeared treating certain aspects of the effects of random irregularities in feeders, but they do not directly apply to the case discussed in this Appendix.

* M. Didlaukis and H. Kaden: *Elektrische Nachrichten-Technik*, 1937, vol. 14, p. 13. P. Mertz and K. W. Pfleger: *Bell System Technical Journal*, 1937, vol. 16, p. 541

[The discussion on this paper will be found on page 138.]

E.M.I. CATHODE-RAY TELEVISION TRANSMISSION TUBES[*]

By J. D. McGEE, M.Sc., Ph.D., and H. G. LUBSZYNSKI, Dr.Ing.[†]

(*Paper first received 20th July, and in revised form 22nd October,* 1938; *read before the* WIRELESS SECTION *7th December,* 1938.)

SUMMARY

This paper gives a short history of the development of two types of cathode-ray tube now used for transmission of television pictures. The construction and mechanism of operation of the "Emitron" is described and its limitations are discussed. A further development, namely the "Super-Emitron," is then described, and its mechanism of operation and its performance are compared with those of the "Emitron." Some other types of transmitting tubes, which have been tested experimentally in the course of this work, are described briefly.

INTRODUCTION

The earliest suggestion that the cathode-ray tube could be used for the transmission of pictures by television was made by Campbell Swinton[‡] in 1908. He realized the limitation of mechanical scanning methods and saw in the then recently-discovered cathode rays a practically inertialess scanning means.[§] He proposed to focus an image of the scene to be transmitted on to a mosaic of photo-electric cells. The elementary cells would become electrically charged due to the liberation of electrons from them by the light, and could be periodically discharged by a beam of electrons which scanned over them. He developed this idea in greater detail in 1911,[||] and in 1926[¶] he attempted to carry it out experimentally, using selenium as a photo-sensitive mosaic in a manner similar to that recently adopted successfully by Miller and Strange.[**] At about this time the number of suggestions for cathode-ray tubes for television picture transmission increases rapidly in the patent literature.[††]

Apart from cathode-ray scanning the most important development in television transmitting tubes is the introduction of charge storage. This paper is chiefly concerned with transmitting tubes employing this principle.

The essence of the principle is that during part at least of the picture-frame period the photo-electric emission due to the light from each picture point of the image is stored up as a charge on a mosaic of photo-electric cells, each associated with a condenser. The condensers are discharged in sequence by some switching mechanism such as a beam of electrons; the sequence of electrical pulses thus produced constitutes the "picture signal." This principle appears to be implied in Campbell Swinton's suggestions. For example, in his presidential address to the Röntgen Society on the 7th November, 1911,[‡‡] he makes the following remarks about the mosaic of photo-electric cells which he proposed to employ in the transmitting tube he suggested: "It is further to be noted that as each of the metallic cubes in the screen acts as an independent photo-electric cell, and is only called upon to act once in a tenth of a second, the arrangement has obvious advantages over other arrangements that have been suggested in which a single photo-electric cell is called upon to produce the many thousands of separate impulses that are required to be transmitted through the line wire per second, a condition which no known form of photo-electric cell will admit of. Again, it may be pointed out that sluggishness on the part of the metallic cubes . . . in acting photo-electrically in no wise interferes with the correct transmission and reproduction of the image, provided all portions of the image are at rest."

Though the storage principle is not mentioned explicitly in this, or any other paragraph, of Campbell Swinton's writings on television, it is difficult to escape the conclusion that it is this principle he had in mind in the first of the paragraphs quoted. This is borne out by the specific and separate reference to the sluggishness of the photo-electric effect in the paragraph immediately following. This interpretation of Campbell Swinton's suggestion is in agreement with that given in 1932 in the excellent handbook on television edited by Prof. Fritz Schröter.[‡]

These early schemes were undoubtedly impracticable at the time they were suggested, and it was only after 20 years of technical progress that it became possible to put them into practice. Furthermore, all the schemes proposed seemed to be theoretically unsound[§] and it is probably true to say that transmitting tubes of this kind were in operation experimentally for a considerable time before the essential feature of their mechanism of signal production was understood.

As far as it has been possible to ascertain, the first disclosure in the patent literature of a photo-electric mosaic electrode is due to Zworykin in 1925,[||] but it seems significant that though the charge-storage principle is used it is not mentioned in the claims. In fact it appears that this fundamental invention was never patented, although in subsequent publications its importance is emphasized.

The first published description of a successfully operated television transmitting tube employing both cathode-ray scanning and charge storage was due to Zworykin[¶] in 1933. Though difficult to manufacture and operate, this transmitting tube, named the "Iconoscope" by Zworykin, is a great advance on the earlier mechanical methods. Only 5 % to 10 % of the possible increase in sensitivity due to charge storage is realized in this tube, nevertheless it is sufficiently sensitive to enable high-definition pictures of studio and outdoor scenes with moderate illumination to be televised.

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[*] Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 468.

[†] Electric and Musical Industries, Ltd.

[‡] See Reference (1).

[§] For an appreciation of the work of Campbell Swinton on television see an article in *Wireless World*, 1935, vol. 8, p. 591; also *Nature*, 1936, vol. 138, p. 674.

[||] See Reference (2).

[¶] *Ibid.*, (3).

[**] *Ibid.*, (4).

[††] *Ibid.*, (5).

[‡‡] *Ibid.*, (2).

[‡] See Reference (6).

[§] *Ibid.*, (6).

[||] *Ibid.*, (7).

[¶] *Ibid.*, (8).

About a year before Zworykin published his description of the Iconoscope, experiments had been begun in the E.M.I. Research Laboratories on the charge-storage type of transmitting tube. These experiments, carried on independently of Zworykin and his co-workers, soon showed that it was possible to generate picture signals in this way. Briefly, the method employed was as follows. An aluminium plate was coated with a thin layer of aluminium oxide by anodic oxidation, and a mosaic of patches of silver was formed on the surface of the oxide by evaporating silver on to the oxide surface through a grid.* This mosaic was mounted as target in a tube, which was then evacuated, and the silver patches were oxidized and activated photo-electrically with caesium. When an optical image was focused on the mosaic while it was being scanned by an electron beam, picture signals corresponding to the optical image were obtained and an image was reproduced on the screen of a cathode-ray tube.

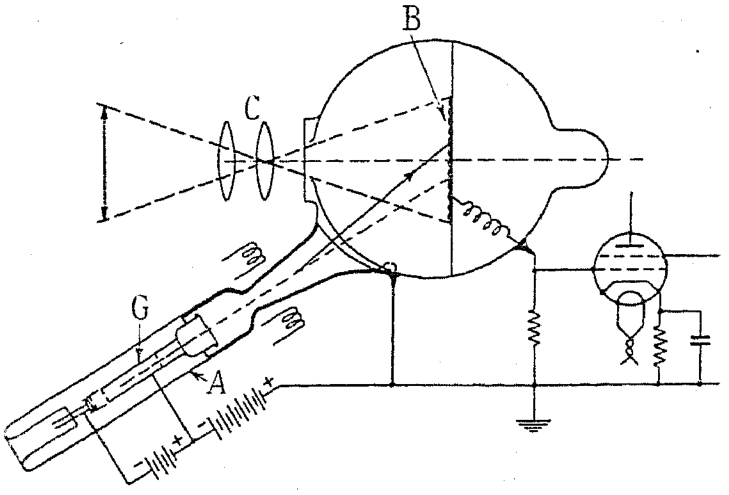

Fig. 1

These experiments led to the development of the Emitron, which is believed to be different in several respects from the Iconoscope and has been in regular service at the London Television Station since its inception in August, 1936. A further development is the " Super-Emitron,"† which was put into service for the first time during the outside broadcast from the Cenotaph in November, 1937. This tube shows a considerable increase in sensitivity and flexibility over the standard Emitron.

THE EMITRON

Construction

The main constructional features of an Emitron are shown in Fig. 1. A spherical bulb of Pyrex glass about $7\frac{1}{2}$ inches in diameter is provided with a neck A in which is fitted an electron gun G which directs a sharply-focused beam of electrons on to the photo-electric mosaic B. A polished flat glass window is sealed on to the bulb as shown, so that an undisturbed optical image of the scene to be transmitted may be focused on the mosaic by means of the lens system C.

The photo-electric mosaic is formed on the surface of a sheet of mica about 0·001 in. thick and 4 in. × 5 in. superficial dimensions. This sheet of mica is carefully selected for uniformity of thickness and freedom from blemishes of any kind. The reverse surface from that on which the mosaic is deposited is coated with a conducting, highly reflecting metallic layer, generally by painting on several coats of commercial " liquid silver." This sheet of mica is supported on a second circular sheet which fits closely into the bulb. The assembly is flexible and can be rolled into a cylinder while being inserted into the bulb, where it is fixed in position.

Electron Gun Design

A diagram of the electron gun used in the Emitron is shown in Fig. 2. The cathode A is of the indirectly-heated oxide-coated type and is placed immediately behind an aperture in a modulating electrode B. The first anode C extends from the modulator to a short distance within the second anode D, which is formed by depositing a layer of silver on the inside of the neck and part of the bulb, as shown.

Since the dimensions of the mosaic which is scanned by the electron beam are approximately 10 × 12·5 cm., the scanning lines being parallel to the longer sides of the rectangle, it follows that for a 405-line picture the scanning spot must ideally be 0·25 mm. in diameter. Since the scanning beam falls on the mosaic at an angle of 35° to the normal, the actual diameter of the electron pencil on reaching the mosaic surface must be kept less than about 0·20 mm. Referring again to Fig. 1, it is clear that the distance travelled by an electron from the main electron lens of the electron gun to the mosaic is greatest when the electron beam is at the top of the mosaic and least when it is at the bottom. Since the electron beam is in focus at one distance only from the electron lens it follows that if it is focused at the centre of the mosaic it will be slightly out of focus at both the top and the bottom, and the reproduced picture will lack detail in these regions. This can be corrected by using in the electron gun an auxiliary electrode which is modulated synchronously with the scanning fields in such a way as to change the focal length of the electron lens slightly—sufficiently to keep the beam in focus over the whole mosaic.*

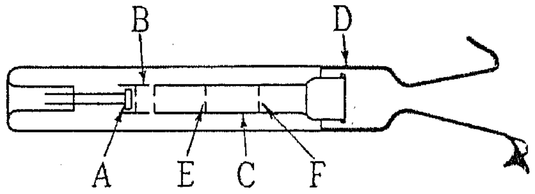

Fig. 2

It is found, however, that this can be achieved more simply by taking advantage of the fact that a comparatively small beam current of the order of 0·10 μA is required for scanning the mosaic. It is therefore possible to reduce the " aperture " of the electron lens to a very much smaller diameter than is possible in a cathode-ray tube for television reception, and consequently very much greater depth of focus of the electron beam can be obtained. Referring to Fig. 2, the aperture E limits the electron beam to a diameter of approximately 2 mm. in the region of the electron lens, which is about 40 mm. in diameter. The aperture F is provided to prevent most of the secondary electrons liberated from the walls of the gun or from the aperture E reaching the mosaic. If such stray secondary electrons reach the mosaic they reduce the efficiency of the tube and produce spurious signals which interfere with the required signals. The fact that a beam of small cross-section suffices has the further advantage that the electron lens need not be carefully corrected, since only a small portion of its cross-section is used.

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* See Reference (9).

† " Super-Emitron " is the trade name which has been adopted for the improved type of transmitting tube described below.

* See Reference (10).

The Mosaic Mounting

As above described, the mosaic is formed on a sheet of mica which is attached to a second sheet serving to anchor it to the bulb. This construction is not completely rigid and in practice it is often necessary to operate tubes under conditions where they are subjected to considerable vibration; for example, on a movable truck in a studio or in a high wind out of doors. It was soon found that such vibration produced spurious microphonic signals which could be divided into two fairly distinct classes, (1) a low-frequency signal at about 100 cycles per sec. and (2) a higher-frequency signal at about 10^4 cycles per sec. The low-frequency microphonics were found to be due to the vibration of the mica disc supporting the signal plate. Any considerable vibration of this kind would have the effect of throwing the optical image rapidly in and out of focus, and hence would result in a blurred picture.

In operation, the signal plate is held at a positive potential relative to the second anode, and the small changes in its capacitance to its surroundings due to vibration set up voltage fluctuations across the signal resistance. This was eliminated sufficiently for all practical purposes by anchoring the supporting mica disc to the glass walls at four points by soft damping wires.

The high-frequency microphonic signals were eventually traced to a similar effect due to movement between the signal plate and the surface of the mica disc on which it is supported. This movement is partly in the plane of the disc and partly normal to it, and the microphonic signals may be produced in two ways. Firstly, friction produces electrical charges between the signal plate and the mica surface. Slight movements of the signal plate normal to the surface of the supporting mica sheet produce changes of the capacitance between these two surfaces. Thus varying potential differences are set up which lead to spurious signals. Secondly, charges may be produced on the surface of the supporting mica sheet owing to electrical leakage over its surface. Any small movement of the surfaces normal to one another would then produce signals. This type of microphonic signal was effectively eliminated by covering the surface of the supporting mica disc, with which the signal plate is in contact, with a conducting metallic coating. This coating is connected electrically with the signal plate and hence held at the same potential.

Emitrons in which these precautions have been taken can be used under conditions where they are subjected to considerable vibration without microphonic signals being generated.*

The Mosaic Surface

The silver mosaic may be formed on the clean surface of the mica by depositing a continuous sheet of silver and causing this to aggregate by heating in air to a temperature of about 700° C.† If the layer is sufficiently thin it breaks up into a series of discrete areas under surface-tension action, even though the temperature be below that at which bulk silver melts. This phenomenon was first observed by Faraday and was investigated more fully by Beilby.‡ The process of aggregation can be clearly seen in the microphotographs shown in **Fig. 3** (see Plate 1, facing page 134);§ these illustrate a series of layers of silver of decreasing thickness each heated to the same temperature. It is seen that when the silver is thick the effect of the heating is to cause a number of holes to appear. As the layer is made thinner these holes increase in size until they merge with one another, leaving islands of silver completely separated. With still thinner layers these islands disintegrate to form still smaller areas.

Exactly the same process is observed in a single silver layer as the temperature of baking is raised from a low value to about 700° C. At comparatively low temperatures (200°–300° C.) the holes appear; as the temperature is raised these increase in size until they merge, leaving the discrete mosaic areas which break up further as the temperature is raised still higher.

That this phenomenon is due to surface tension in the layer may be seen by comparing these photographs with those obtained by Beilby‖ for the aggregation of oil films under surface-tension forces.

Although photo-electric mosaics formed in this way are in many ways satisfactory, there are a number of disadvantages such as the difficulty of ensuring that the mosaic, which must be formed before being sealed into the tube, is kept clean and free from spots and other blemishes.

A more satisfactory way of forming the mosaic will be described in a later communication by L. Klatzow.

Mechanism of Signal Production¶

When a photo-electric mosaic of the type described above is scanned by an electron beam of about 1 000 volts velocity, each primary electron liberates several secondary electrons from the surface. The behaviour of an insulated conducting target of mosaic character when bombarded by electrons may be shortly recapitulated. Fig. 4 shows a typical curve of the number of secondary electrons released per primary electron as a function of the velocity of the primary electrons. The curve possesses a maximum, the position of which, for most substances, lies between 300 and 400 electron volts. There are two values of the velocity of the primary electrons for which the secondary-emission coefficient is unity, viz. those represented by the points A and B in Fig. 4. We may call A and B the "first and second cross-over points" respectively. Under the influence of the primary beam and the loss of secondaries the target takes up a certain equilibrium potential. There are three different points on the characteristic for which the potential of the target will be stable.

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* See Reference (11).

† *Ibid.*, (12).

‡ *Ibid.*, (13).

§ The authors are indebted to Dr. L. Klatzow, of the E.M.I. Research Department, for these photographs.

‖ See Reference (13).

¶ The mechanism of signal production in the Iconoscope was recently discussed by Zworykin, Morton, and Flory (see *Proceedings of the Institute of Radio Engineers*, 1937, vol. 25, p. 1071).

First, if the primaries hit the target with a velocity below that corresponding to A, they will charge the target negatively, until it stabilizes at the potential of the cathode from which the primary electrons emerge, i.e. at the potential represented by the point O on the curve.

The second possibility occurs when the velocity is between those corresponding to the points A and B. In this case the secondary-emission coefficient is greater than unity, and equilibrium will be attained when, for each primary electron striking the target, one secondary electron (of the number released) reaches the collecting electrode. This equilibrium potential depends mostly upon the geometrical configuration of the tube and the secondary-emission coefficient. The equilibrium potential of the target will be in the neighbourhood of that of the collecting anode—a few volts above or below.

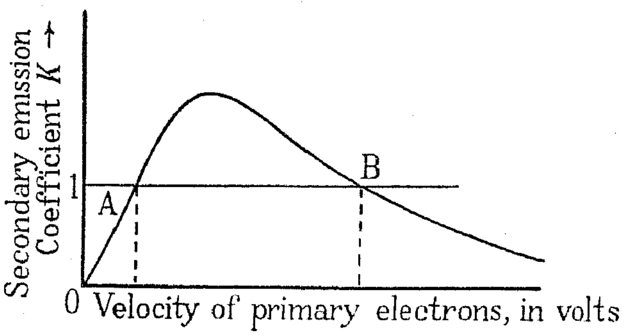

Fig. 4

The third possibility occurs when the primary electrons reach the target with a velocity greater than that corresponding to the second cross-over point B. In this case the number of secondaries released will be less than the number of primaries which strike the target, and the target potential will, therefore, become more negative until the primaries are decelerated to a velocity corresponding to the second cross-over point. When this point is reached one secondary electron will be released per primary and the potential of the target will be stable. In this case the equilibrium potential is independent of the potential of the collecting anode but has a value fixed with respect to that of the cathode. It is thus possible to set up a potential difference between the target and the collecting electrode.*

The problem is complicated when a target consisting of an insulator or of a mosaic of minute conducting elements deposited on an insulator is scanned by a fine pencil of electrons. This is the condition which occurs in the Emitron. The potential across the surface of the target will no longer be uniform but will vary from point to point, and from instant to instant as the scanning beam explores the surface. The average potential of the target at any instant will be determined by the secondary-emission coefficient of the surface and the geometry of the electrodes. For example, at a velocity of the primary electrons for which the ratio of secondaries to primaries is greater than 1, the potential will be such that on an average during at least one scan the number of secondaries arriving at the collecting electrode will be equal to the number of primaries which reach the mosaic. In the Emitron, where the collecting electrode is the second anode of the gun, it has been found that the average potential of the mosaic when scanned in the dark is about 1·5 volt negative with respect to the collecting anode. The value obtained by Zworykin* and his co-workers for the Iconoscope is in agreement with this figure.

On the other hand, the potential of an elementary area on the target of the size of the scanning spot will vary periodically as the beam scans the surface. Each primary electron liberates K secondaries which have a velocity distribution varying from almost zero to the velocity of the primary electrons, but the great majority of these secondaries have velocities of the order of 2 to 4 volts. It is clear that, if the point on the mosaic on which the beam impinges is initially uncharged, it will at first lose practically all the secondary electrons liberated from its surface, and consequently will begin to rise rapidly to a more positive potential. As it rises in potential the slower secondary electrons will find it impossible to escape to the surrounding parts of the mosaic, and finally the element will reach a potential such that only one secondary electron can escape per primary arriving. It is found that in this state the elements immediately under the electron beam are about 4 volts positive with respect to the surrounding parts of the mosaic. This rise in potential of a scanned mosaic element is shown in Fig. 5.

When the scanning beam has passed an element the latter will collect secondary electrons from other elements being scanned, and its potential will soon return to approximately the average potential.

It follows that there is quite a strong electric field between the mosaic element which has reached this positive potential under the scanning beam and the surrounding elements. Hence any photo-electrons liberated by the light from the elements that are about to be scanned will be collected by the elements that are being or have just recently been scanned, and the electric field is sufficient to saturate all such photo-electrons up to a distance of the order of 1 cm. in front of the scanning beam. Thus the scanning beam builds up a positive potential front as it scans over the mosaic, which in turn builds up a distribution of electrostatic charges in front of it, corresponding to the light distribution in the image, thus giving very efficient charge storage over a small fraction of the total frame period.

This phenomenon was called "line sensitivity" by Zworykin. In the authors' opinion the Emitron operates almost entirely on this line sensitivity. The integrating sensitivity, i.e. that part of the signal which is due to storage over the whole of the frame time, is comparatively low. This can be seen from the following observations.

A fast-moving object which moves in a direction parallel to the scanning lines is imaged on the receiving screen as a series of pictures, each of which is perfectly sharp, with no signal between the single pictures. This is due to the mosaic elements becoming substantially

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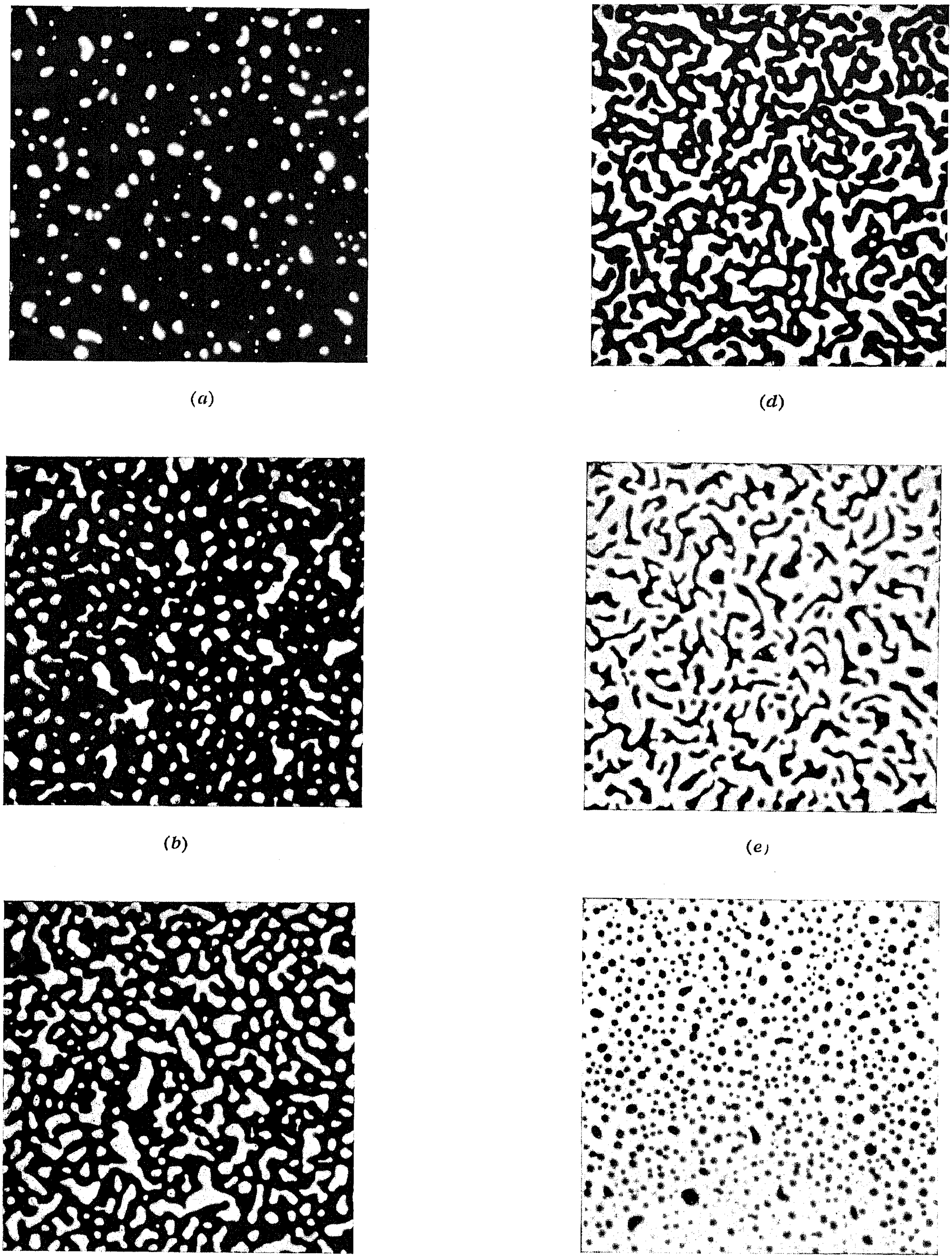
* See Reference (14).

* See Reference (15).



(*a*)

(*d*)

(*b*)

(*e*)

(*c*)

(*f*)

Fig. 3

Plate 2

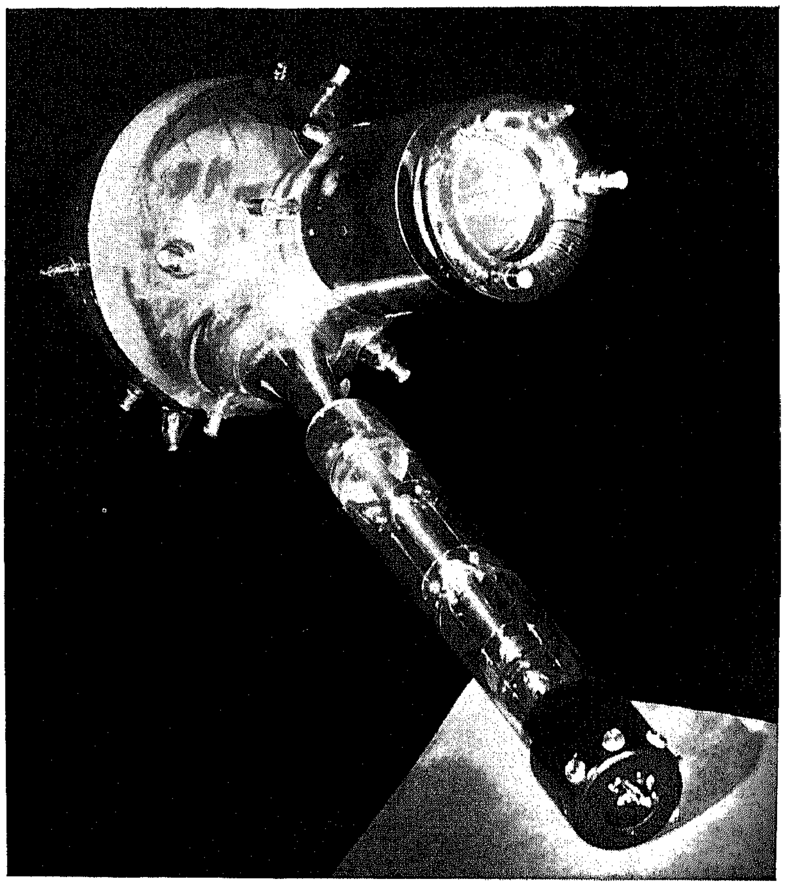


Fig. 7

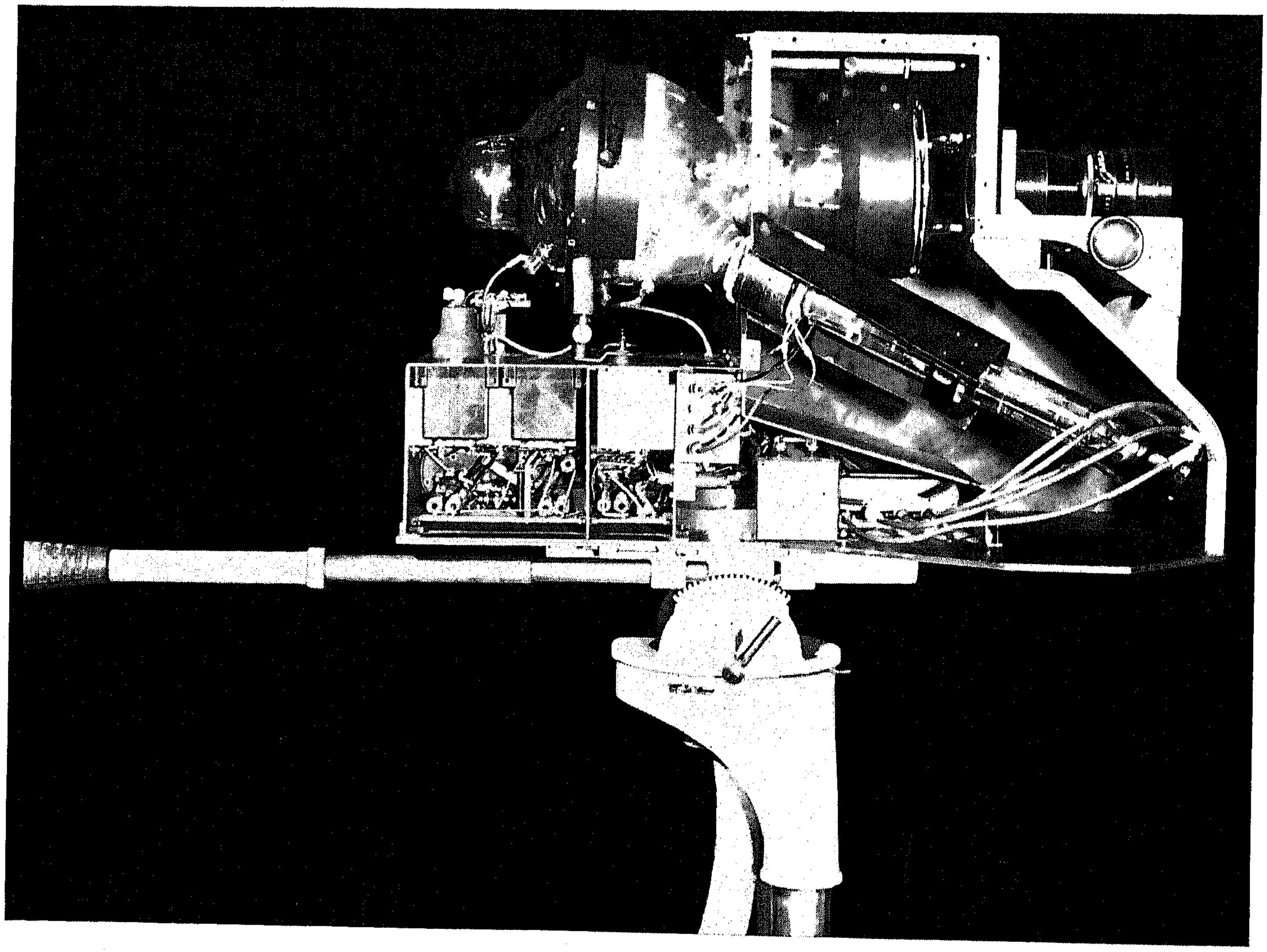


Fig. 8

sensitive only immediately before the scanning beam reaches them. If an integrating sensitivity were present to any appreciable degree, the picture of the object would have a sharp leading edge with a blurred trail behind extending to the leading edge of the next picture.

In transmitting pictures from cinematograph film it is found that if the image is projected continuously on to the mosaic while the electron beam is scanning the mosaic, very much smaller light intensity is required than if the separate pictures are projected on to the mosaic during the intervals between frames when the electron beam is suppressed. Even when allowance is made for the difference in the time that the mosaic is exposed to the light image, the efficiency of the tube appears to be of the order of 10 times greater in the former case.

Finally, the result of applying a positive potential to the collecting electrode leads to the same conclusion. It can be seen by reference to Fig. 5 that the actual picture signal is derived from the differences in the potential changes which occur when a dark and when an illuminated element is scanned by the beam. If the potential of the second anode, which acts as the collecting electrode in the Emitron, is suddenly raised to a positive potential with respect to the signal plate which is sufficient to ensure saturation of photo-electrons liberated from the mosaic, it is found that a strong signal is obtained, which slowly dies away. Now a signal can only be obtained if the stream of secondary electrons leaving the mosaic varies according to the distribution of light in the optical image. Hence, the secondary electron emission cannot be saturated, since then no signal would be produced, although the potential between the second anode and mosaic is sufficient to saturate slow electrons liberated from the mosaic in the absence of the scanning beam. It follows that the flow of secondary electrons from the mosaic to the second anode must be controlled by the potentials of the mosaic elements in the immediate neighbourhood of the scanning spot, rather than by the potential of the second anode, since a potential difference of 3 or 4 volts between the element which has just been scanned and the element which is just about to be scanned requires a field of the order of 100 volts/cm., while the field due to the saturating potential on the second anode is only of the order of 10 volts/cm. at the mosaic surface.

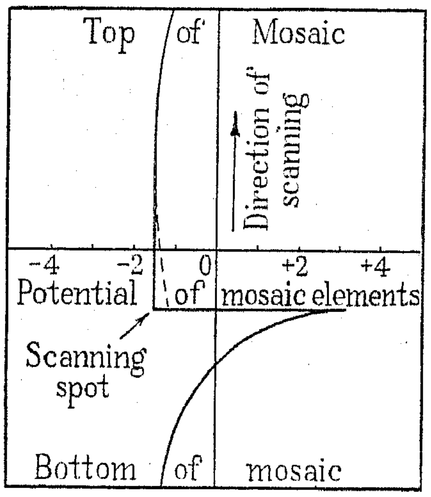


Fig. 5

The application of the saturating potential will, however, result in almost all photo-electrons which are liberated by light at points of the mosaic some distance from the scanning spot being collected by the second anode. Hence, a stronger electrostatic image will be built up than is obtained in the normal working of the tube, and the signals will be proportionately stronger. Because of the saturating field the proportion of secondary electrons that escape from the mosaic will be slightly increased, and the average potential of the mosaic will drift towards the potential of the second anode. Eventually it will reach a new equilibrium potential of a few volts negative with respect to the second anode, and the signals will return to normal. The time taken for the mosaic to reach its new equilibrium potential depends on the scanning beam current and the capacitance of the mosaic elements to the signal plate.

If the potential of the collecting anode could be raised to a sufficiently high value to saturate all the secondary as well as photo-electrons, the number of electrons leaving the mosaic would be independent of the illumination of the scanned element, and consequently no picture signal would be produced. It has not been possible to apply a sufficiently high potential between the second anode and signal plate to realize this state completely, but an applied potential of 750 volts was found to produce initially a weak signal, which increased considerably as the mosaic potential drifted towards second anode potential, passed through a maximum, and then returned to normal.

THE SUPER-EMITRON

Theory of Operation

As described above, the Emitron operates with a low efficiency, of the order of 5 % of the theoretical maximum. The reason for this low efficiency is the lack of saturation of the photo-emission from the mosaic during most of the frame period. Another reason, though probably not so important, is the spread of secondary electrons released by the scanning beam from the mosaic. These secondaries neutralize the charges stored on the mosaic elements and also generate spurious signals, as, for instance, a low-frequency component known as " tilt " which is superimposed on the picture signals. The tilt may be corrected by suitable electrical circuits.*

It was considered† that by separating the two functions of photo-emission and charge storage it might be possible to improve upon the efficiency of the Emitron.

A tube which enables this suggestion to be carried into practice is shown diagrammatically in Fig. 6. The optical picture to be transmitted is focused on a continuous transparent photo-surface P. An electron beam is generated at this surface which varies in density across its cross-section according to the illumination. This electron beam is accelerated by the electrode A_2 and focused electron-optically by means of the magnetic lens L on to the storing mosaic M, which is backed by a signal plate S. The mosaic is not photo-sensitive. The beam of photo-electrons projected on to the mosaic generates there a charge distribution which corresponds to the optical picture on the photo-cathode P. The mosaic is scanned in the usual manner by an electron beam from a gun G, thus restoring the elements to an equilibrium potential.

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* See Reference (16).

† *Ibid.*, (17).

There are three striking advantages in the new tube; (1) the improvement in the efficiency of the photo-cathode, (2) the multiplication of the charges on the mosaic by secondary emission, (3) and the improved optical conditions.

In the normal Emitron the photo-sensitivity is limited to about 12 μA/lumen. Continuous photo-cathodes can at present be made with photo-electric sensitivities more than twice that obtained from mosaics in the Emitrons. There are several reasons for this. For one thing, the ratio of the surface covered by photo-electric material in a mosaic to the total area is only of the order of 50 %. Furthermore, the Emitron mosaic has to carry out two functions—it has to emit photo-electrons and to store charges. For storing charges efficiently the insulation between the mosaic elements must be high, and the quantity of caesium which is applied in forming the photo-mosaic is therefore limited as a compromise between sensitivity and insulation.

The other main feature of the new tube is the secondary amplification at the mosaic surface. Under the impact of each photo-electron from the cathode P a number of secondary electrons are liberated at the mosaic and the elements are left with a net positive charge. The sign of the charge stored on the mosaic is the same as in the Emitron, and the signals therefore have the same sense. The amount of charge stored is multiplied according to the secondary emission-coefficient K, the multiplication factor being $(K - 1)$. For caesiated surfaces the factor K is of the order of 7 to 9 for an optimum primary velocity of the electrons of about 400 volts. Therefore, under the same condition and for equal photo-sensitivities it is to be expected that the signal from the new tube should be 6 to 8 times stronger than that from the standard tube.

Comparative measurements have been made on the signal output of a number of Emitrons and Super-Emitrons for equal conditions of illumination and beam current. These measurements showed that for equal photo-sensitivities the new tube gave about 10 to 15 times the signal amplitude of the Emitron. In a few cases the ratio was considerably higher.

This factor, therefore, is too large to be explained in terms of secondary-emission amplification only, but is satisfactorily explained if account is taken of the different velocities of the photo-electric and secondary electrons. As was shown above, the mosaic elements are brought to an equilibrium potential by the scanning beam, which potential is very close to that of the second anode of the gun. The integrating efficiency of the Emitron is small compared with the "line sensitivity," since the photo-electrons can only escape from the elements a short time before the scanning beam strikes them. The photo-electrons released by visible light from the photo-sensitive mosaic have velocities corresponding to approximately 1 electron volt. In the new tube, secondary electrons are released from the mosaic elements. The bulk of these have velocities of several electron volts, and a certain proportion have much higher velocities. They will, therefore, escape more readily under the same field conditions and reach the second anode. This will result in an increase of the integrating efficiency, since a substantial number of electrons will be able to leave the elements between scans. This is confirmed by the observation that fast-moving objects imaged by the new tube leave a blurred trail behind them in the picture, the leading edge being sharp.

Construction

A diagram and a photograph of this type of tube as now in production are shown in Fig. 6 and in Fig. 7 (see Plate 2) respectively.

The electron gun and the mosaic are arranged in the same relative positions as in the Emitron. The chief difference is the addition of the projection neck for the electron optical conversion of the light picture into the electron picture. At the end of this neck is an optically flat window in front of which the photo-cathode is fixed. It is of the transparent type formed on a mica disc approximately 60 mm. in diameter. The cathode is surrounded by a short cylinder for concentrating the photo-electron beam. The second anode of the scanning gun is extended into the projection neck and serves to accelerate the photo-electrons. It is formed by depositing a silver coating on the walls of the tube. The beam of electrons from the photo-cathode is focused on the mosaic by a magnetic lens L of the ironclad type. The coil, however, also produces a rotation of the picture about its axis, and, in order to correct for this rotation, the tube is rotated in the camera through the same angle, and in the same sense. A photograph of the tube in the camera is shown in Fig. 8 (see Plate 2); the focusing coil and the rotation of the tube in the camera are clearly visible. The electron optical magnification between mosaic and photo-cell is approximately 4 : 1. The mosaic consists either of an aggregated aluminium layer on mica or of plain mica which is coated on the back with the usual conductive coating to form the signal plate.

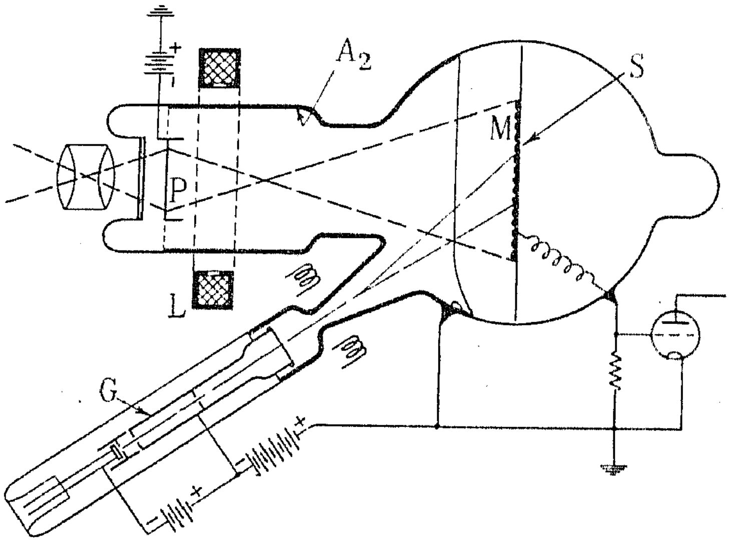

Fig. 6

As the size of the optical picture on the photo-cathode is much smaller than that on the mosaic of the Emitron, lenses with shorter focal lengths are used. Furthermore, owing to the increased sensitivity of the new tube, smaller lens apertures can be used for the same illumination resulting in greatly increased focal depth of the transmitted pictures.

Careful screening in the camera is necessary to prevent interference between scanning and focusing magnetic fields.

When used in this way the new type of tube gives very satisfactory results and has been frequently used by the B.B.C. when the conditions are such that very little light is available or where telephoto lenses are necessary.

The sensitivity of the new type of tube can be still further increased if the electron beam forming the picture is amplified by secondary emission one or more times before reaching the mosaic.* Development along these lines is being carried out and shows considerable promise.

OTHER FORMS OF CATHODE-RAY TUBE

A number of other forms of television tubes which may be mentioned here have been constructed and used experimentally.

The type of tube described above may be used under different operating conditions from those mentioned.† Referring to Fig. 4, if the photo-electrons are accelerated sufficiently to strike the mosaic with a velocity much beyond that corresponding to the second cross-over point B, they will release only a few secondaries and will charge the elements negative. The scanning beam strikes the mosaic with a velocity corresponding to a point between A and B and restores them to an equilibrium potential. A tube operated in this way produced signals the sense of which was opposite to those produced by the standard tube. This method, however, does not give secondary amplification.

Another method is to scan the mosaic with a beam the velocity of which is greater than that corresponding to the second cross-over point.‡ A potential difference is then established between the mosaic and the second anode, which is equal to the difference between the potential of the latter and that of the second cross-over point. If this difference be made large enough, the whole of the emission from the mosaic will be saturated and collected on the second anode. There will be no spread of secondary electrons and therefore no tilt. Furthermore, there will be the full theoretical efficiency due to the complete saturation of all the electrons from the mosaic. The photo-electrons are projected on to the mosaic with a velocity corresponding to the peak of the secondary-emission characteristic in Fig. 4. Those elements which are hit by photo-electrons in the intervals between scans rise in potential and the scanning beam restores them to their equilibrium potential at the second cross-over point. The sense of the signal is the same as that of the standard tube.

As the scanning beam releases very nearly 1 secondary electron per primary, its impedance is very high and consequently large scanning-beam currents must be used in order to restore the elements completely to their equilibrium potential during a single scan. It can easily be shown that an element is restored by the beam to its equilibrium potential exponentially with a time-constant $T = C/(i_p a)$, where C is the element-to-signal plate capacitance, i_p is the primary current in the scanning beam, and a is the slope of the secondary-emission characteristic at the second cross-over point. To ensure complete discharge the time-constant must not exceed about one-third of the time the beam remains on the element. It is necessary, therefore, in order to keep the beam current small, to reduce the element to signal-plate capacitance and to choose a material for the mosaic for which the slope of the secondary-emission characteristic at the second cross-over is as large as possible.

Carbon was found to be the most suitable material for this purpose, and a tube with a carbon mosaic was successfully worked in this way. A very much increased sensitivity and a complete absence of "tilt" were obtained with scanning-beam currents of 55 μA, whereas normally the beam current must be limited to about 0·1 μA to reduce "tilt" to manageable dimensions.

Other forms of tube have been constructed and used in which the necessity for scanning the mosaic at an angle is avoided, and it is therefore unnecessary to supply circuits to correct for trapezium distortion. In the normal Emitron this was done by making the signal plate transparent* and projecting the image on to the mosaic through the signal plate. The sensitivity of such a tube is, however, comparatively low.

Finally, trapezium distortion may also be avoided by using a "double sided" mosaic. In these mosaics the elements are in the form of insulated rivets in a mesh grid which acts as signal plate.† The rivets project on both sides of the grid and may be photo-sensitized on one side. The image is focused on this side and the scanning beam explores the reverse side, the axis of the gun being normal to the plane of the grid. Such double-sided mosaics may also be used in the type of tube where the electron image is focused electron-optically upon the mosaic.‡ In this case it is not necessary to photo-sensitize the rivets.

Acknowledgments

Acknowledgments are due to Mr. I. Shoenberg, the Director of Research; to Mr. G. E. Condliffe; and to the staff of the Research Department, among whom Messrs. L. Klatzow, J. E. I. Cairns, and H. Miller, should be mentioned.

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DISCUSSION BEFORE THE WIRELESS SECTION, 7TH DECEMBER, 1938, ON THE PAPERS BY MESSRS. CORK AND PAWSEY (SEE PAGE 111) AND MESSRS. McGEE AND LUBSZYNSKI (SEE PAGE 131)

Mr. L. H. Bedford: With regard to the paper by Messrs. Cork and Pawsey, it is a matter for some surprise that so high a degree of adjustment and refinement in regard to feeders and terminations is either necessary or possible. It would seem, however, that the accuracy of the results given may be somewhat artificial. On the one hand, we are working in zones where the impedance of 1 in. of wire is very high relatively to the impedance changes concerned; on the other hand, it may be questioned whether the representation of a feeder as a distributed inductance and capacitance can be pushed this far, bearing in mind that this representation is after all only an approximate form of a complicated field theory. The results, however, appear to be adequately justified in practice, not only as shown by the authors' curves but also as seen by any viewer whose receiver does not itself introduce more frequency distortion than that with which the authors are concerned.

The paper by Messrs. McGee and Lubszynski considerably clarifies the relation between the Emitron and the Iconoscope. It would appear from the data given that the distinction is perhaps more one of history than of function.

The great advance which has been made in transmission in the last 2 years is largely attributable to improvement in the Emitron camera, but there is still a shortcoming in connection with film transmission. In my opinion the use of a Mechau projector is not an entirely elegant solution to the problem, and in practice there are evidently great difficulties of adjustment, as is evidenced by the frequent occurrence of flicker or differential illumination of the interlaced rasters. The original method employing intermittent film movement and making use of the "memory" of the Emitron was theoretically more elegant, and failed only because of the imperfect memory characteristics of the Emitron. The paper clarifies this phenomenon and would indicate that with developments along the lines of the Super-Emitron a reversion to the original film method might be contemplated.

Mr. D. C. Birkinshaw: At the time of installation of the feeder Messrs. Cork and Pawsey were apprehensive of the effect of weather upon it; they thought that damp conditions might lower the feeder resistance, and possibly introduce capacitance variation. Measurements made over the last 2 years, however, show that the resistance of the feeder is more than 10 megohms in dry weather, and never descends below about 30 000 ohms in wet conditions. There is also no evidence that weather conditions introduce any capacitance variations. Although means have been provided for electrically drying out the feeder if necessary, it has so far been unnecessary to bring them into operation.

One interesting feature of the feeder which is not mentioned in the paper is the fact that it was unfortunately impossible in one case to obtain a pipe of the correct diameter for the inner conductor in connection with the first transformation to 78·5 ohms, and it was necessary, therefore, to transform to an impedance somewhat greater than this figure, and insert a further transforming section to correct for the error.

Turning to the paper by Messrs. McGee and Lubszynski, I should like to mention that in the early days of the Emitron its colour response was characterized by an emphasis of the red and infra-red end of the spectrum. This introduced certain make-up problems, and gave the picture a peculiar appearance similar to that of a photograph taken on an infra-red-sensitive plate. This was undesirable, and was later overcome, and the tubes now have their peak response in the green region. This considerably improves the quality of the picture, and lessens the difficulty of the make-up problem. I should like to ask the authors how this change in colour response was effected.

Mr. D. C. Espley: Messrs. Cork and Pawsey state that the feeders work from a transmitter termination of 50 ohms to an aerial impedance of about 20 ohms. Is it not possible to use a feeder of lower characteristic impedance? It seems to me that the only limitation would be the voltage on the feeder (about 1 100 volts at present). The change could be made by increasing the diameter of the inner conductor, or preferably by reducing the outer, and in this connection it would be interesting to know the significance of the radius ratio of 3·6. I cannot see why it should be so important to keep to the optimum attenuation figure resulting from this ratio, because the loss in a 450-ft. feeder of that kind works out at less than 0·2 db.

I should be interested to know whether in the course of these precise impedance measurements any effects due to the proximity of the resonant systems of the sound aerial were noticed. In experiments in this general field we have had trouble with sliding joints, as they can introduce appreciable series resistance. Have the authors also noticed such effects?

It would be interesting to have details of the steady-state propagation characteristics of the transmitter, expressed as the relation of the current in the output stage to the current in the aerial. Further, would the order of accuracy demanded of the authors' impedance measurements show up transmission irregularities equally well?

Turning to the paper by Messrs. McGee and Lubszynski, was the lens correction for focus on the mosaic applied along one dimension of the picture, or was it a two-dimensional correction? I notice that the camera is tilted through a certain angle in order to compensate for the image rotation mentioned by the authors, but would not it be possible to get over that effect equally well, and perhaps in a more convenient manner mechanically, by rotation of the deflection field? Such rotation would not have any effect on the depth of focus, which is already taken care of by the use of a very small beam current.

There is one last point to which I should like to refer. In observing received pictures it occasionally seems that the camera has a spot of low sensitivity in the middle of the field. I suppose that this could be due to one of two causes—either to a simple electron burn, or to negative-ion poisoning. Is any device provided for obviating either of these troubles?

Mr. A. H. Mumford: I should like to know what factor determined the size of feeder adopted by Messrs. Cork and Pawsey. Was it the permissible eccentricity of the inner conductor? The principal causes of reflections appear to be impedance irregularities at connecting boxes, and it is stated that these were compensated for by means of variable capacitors. It is not quite clear whether it was necessary to introduce other variable capacitors in the straight portion of the line to compensate the effects of varying eccentricity of the inner conductor. I notice that the maximum eccentricity which the authors suggest should be allowed is 0·05 in., whereas for the Alexandra Palace feeders the figure was 0·15 in. Have they any scheme in mind for reducing the magnitude of this variation in any future stations which they may be called upon to put up?

It would have been interesting if the paper had given some more information as to the transmitter and the aerial circuits. I notice that although the authors find an impedance match with artificial loads of ± 0·5 ohm between 43 and 47 Mc./sec., as soon as they introduce the aerial system the figure goes up to ± 5 ohms. This suggests that possibly the next advance should be in the direction of producing an aerial which has a lower impedance variation. Can the authors say whether the limits to which the aerial and transmission line were adjusted initially have been maintained after some months of operation? It would be interesting to know whether frequent checks of this are made by the maintenance staff, and, if so, the extent of the variations observed.

Turning now to the paper by Messrs. McGee and Lubszynski, considerable attention is paid to the sensitivity of the television-transmission tubes, but obviously the signal/noise ratio obtained from the tubes is also of considerable importance and more information on this point would be very valuable. In particular, is the signal/noise ratio of the Emitron better than that of the Super-Emitron or vice versa? Is any improvement in signal/noise ratio considered possible? On page 137 it is pointed out that the new type of tube gives very satisfactory results and has frequently been used by the B.B.C. If it is so outstandingly good, why has it not been used continuously? I should have considered that its increased sensitivity would have been as valuable for studio transmissions as for outdoor work. Finally, it would be interesting if the authors could give some idea of the life and cost of the tubes described in their paper.

Dr. D. Gabor: I cannot quite accept the claim that Campbell Swinton was thinking of charge storage when he wrote the passage quoted by Messrs. McGee and Lubszynski. There is no doubt that he had only photoconductivity in mind, otherwise he would not have referred to the sluggishness of the cells. This effect is eliminated by the principle of using a great number of cells and scanning them in succession. But as regards storage, a photo-conducting cell is a rather different matter from a photo-electric layer backed by an insulator. The photocell integrates over any length of time. On the other hand, the best a photo-conducting cell can do is to reach an equilibrium with the illumination, and this equilibrium is only delayed by the sluggishness due to transient polarization. Miller and Strange have found certain charge-storage effects, similar to those exhibited by every imperfect dielectric, but these are not due to the illumination but to charges produced by the scanning beam. These produce a "memory" effect, which is, however, pronounced not in the bright but in the dark parts of the picture. The illumination itself produces no surface charges.

May I ask Messrs. McGee and Lubszynski whether in the Super-Emitron any measures have been taken or found necessary to eliminate the pincushion distortion which arises in image convertors with electrostatic lenses, or whether the magnetic lens used was sufficiently free from this defect.

At the end of the paper a very interesting tube is described which operates at the second unity point of the secondary-emission curve. It is stated that carbon is a suitable material for the mosaic, as the curve for carbon has a steep slope near B (Fig. 4). It appears, however, from the experiments of Knoll and his collaborators, and I have also found in my own experiments, that the secondary emission of graphite never reaches unity. Was the carbon in question specially prepared?

Dr. R. F. J. Jarvis: The analysis which Messrs. Cork and Pawsey have made in Section (2) of their paper illustrates the primary importance of time response rather than frequency response in determining the behaviour of a television circuit. It will be noted that the procedure which they adopted was to consider the response of the line and aerial circuit to a sudden change of carrier input voltage, and that by laying down limitations to the form of this response they estimated to a first approximation the greatest unobjectionable impedance mismatch in the frequency range 43–47 Mc./sec.

It is interesting to recall that in the early days of electrical communications, when telegraph signals were the only type of signal transmitted, scientists considered the time response of the circuit in analysing its behaviour mathematically. The Kelvin arrival curve for the submarine cable is a good example of this. With the coming of telephony and music transmission the frequency response of a circuit became more important than the time response, for the following reasons: (*a*) Most speech and musical sounds are quasi-periodic in nature. (*b*) The most important attribute of the ear is its ability to discriminate between sounds of different frequencies. It has been suggested in recent years that the ear is enabled to do this by means of the basilar membrane of the cochlea, with its fibres of different lengths stretched radially like a harp. (*c*) It has been found much more convenient to design complicated circuits on the basis of frequency response rather than time response.

It was also found that the ear was unable to differentiate between the phase of two waves of sound of the same frequency, so that the phase response of a sound transmission circuit was not very important, except possibly in the reproduction of transient sounds.

When we come to television, however, this state of affairs is completely changed. The most important attribute of the eye is its ability to discriminate between the luminous flux density of the light arriving at the eye at different angles to the optical axis. The eye is enabled to do this by the 5 million rods and cones on the retina.

The spot of light forming a television picture moves with uniform velocity, so that its position is a linear function of time. The eye cannot follow the spot, but by reason of the persistence of vision sees at any instant the brightness of the spot at all its positions on the picture over the preceding period of about $\frac{1}{25}$ sec. In this way the variation of the brightness of the spot with time is translated into a variation of brightness in two dimensions in space, thus forming a picture. Thus, although the eye does not see brightness so much as a function of time as of space, the latter function is directly related to time in a picture formed by constant-velocity scanning. The time response of the circuit, which determines the variation of the brightness of the scanning spot with time, is thus the most fundamental in a television circuit.

It is well known that given the time response of a circuit its frequency response can be determined by Fourier analysis, it being remembered, of course, that frequency response is a complex function of frequency. The frequency response of a television circuit must be regarded, however, as a mathematical method of expressing the response not directly related to the physical conditions of the problem. It is true, of course, that the signal produced by scanning a fixed picture is periodic with a fundamental frequency of 25 cycles per sec., but this does not alter the fact that the eye is primarily aware of the time response of the circuit and not the frequency response.

Frequency response is still of great value in analysing mathematically the behaviour of different circuits and relating this to measurements on the circuit, as shown by the present paper, but the point which I think needs to be stressed is that in order to achieve a given quality of transmission we must stipulate, at first, not the frequency response but the time response of the circuit, as is done by the authors.

The conditions which must be applied to the time response of a television circuit must be based fundamentally on the maximum rate of change of spot brightness required to achieve a given definition, and on the minimum perceptible percentage difference of brightness, or perhaps, in some cases, the maximum tolerable percentage difference of brightness in regions which should be of uniform brightness. For the minimum perceptible percentage difference of brightness, I am interested to note that the authors suggest a value of 2 % of the maximum brightness. The minimum perceptible percentage difference of brightness is, according to Fechner's law, nearly constant over a wide range of brightness and equal to about 1 % to 2 %. We find that these figures are confirmed by tests of the interfering effect of additional unwanted signals.

Referring to Fig. 33, I note that the impedance/frequency characteristic of the aerial-feeder system showed a variation of ± 5 ohms in 78 ohms over the frequency range 43 to 47 Mc./sec. As is explained by the authors in Section (2), this means that the radiated energy will vary over a range of ± 0·5 db. for constant input voltage to the output stage of the transmitter over this frequency range. The overall fidelity of the transmitter over the *video* frequency range 0 to 2 Mc./sec. will therefore probably vary over about the same range, assuming perfect response in the modulator over this range. I should be interested to know whether any tests have been made to confirm this.

Mr. T. H. Bridgewater: My few remarks will be confined to the paper by Messrs. McGee and Lubszynski.

The Emitron is a very remarkable device because it seems to have been developed on the basis of the frame storage theory, and then to have been found to work very well but according to quite a different explanation, namely the line sensitivity theory. Perhaps I may be permitted to make a few remarks about the working conditions and behaviour of the Emitron at Alexandra Palace.

The beam current which is generally employed is between 0·05 and 0·15 μA. The spectral response of the mosaic appears to correspond very closely to that of the human eye, such variations as occur between one Emitron and another, or during the life of a particular Emitron, being found towards the red end of the spectrum.

For studio work, the mosaic generally has an illumination of between about 0·75 and 1·5 foot-candles. This seems to be satisfactory for most subjects, except for certain kinds having large black areas, particularly in the lower portion and extending to the borders. When these black areas occur it is sometimes necessary to increase the illumination, particularly on the dark parts mentioned, in order to overcome distortion which shows itself as a rather intense white edging to the black parts, encroaching to some extent beyond the edges of these.

Another form of distortion which is noticed from time to time, generally called "streaking," is the appearance of a white overshoot from a horizontal black area on a light background, or a black overshoot from a white area on a dark background. This effect apparently occurs with all Emitrons, and varies in amount according to the contrast ratio, the degree of illumination, and the beam current. It can be corrected for electrically, provided no variations occur of the controlling factors mentioned. Another common effect consists of a darkening of the top of the image which extends for 10 % or more of the picture height, and varies as between one Emitron and another and during the life of a particular Emitron. Perhaps the authors would explain in a little more detail the causes of these distortions.

Mr. E. K. Sandeman: I notice that Messrs. Cork and Pawsey found it necessary to use a thermostat to control the temperature of the terminating resistance; what steps did they take to control the resistances and impedances used in the substitution method which they adopted? This substitution method is rather novel; was its use decided upon because the authors found difficulty in connection with the conventional bridge method at such a short wavelength?

The authors state that a deviation of $\frac{1}{20}$ in. in the centre conductor induces an impedance variation of 9 ohms, which they consider too much. Apparently they eliminate such variations by rotating parts of the centre conductor which are out of truth, and by various other means. It does not appear to be hard to make a mechanical device true to $\frac{1}{20}$ in., and I should like to know whether any particular difficulties have been experienced in doing this.

Referring to the paper by Messrs. McGee and Lubszynski, it would be interesting if the authors could add a diagram illustrating the variation of sensitivity of the various types of Emitron throughout both the luminous spectrum and the infra-red region.

Mr. W. H. Ward: When television becomes one of our major industries it will be necessary to draw up British Standard Specifications for television cables, and before this can be done a standard test will have to be evolved. As this test will have to be applied by men who are not at present familiar with high-frequency technique the committee which has to draw it up will have a difficult task.

I gather from Mr. Birkinshaw's remarks that the Alexandra Palace feeders vary considerably in insulation resistance with the weather but that their behaviour in their working conditions remains the same. It is possible that this may be explained by the large variation, with humidity, of the power factor of steatite at low frequencies as compared with high.

The depth of penetration of a current at frequencies such as that of the Alexandra Palace transmitter is quite small, and therefore any dirt or copper oxide on the surface of the copper conductors might have an appreciable influence on their effective resistance, and consequently on the attenuation.

I notice that Messrs. Cork and Pawsey have succeeded in using $\frac{1}{2}$-watt metallized resistors as sub-standards. Does it just happen that resistances of about the values which are wanted are nearly independent of frequency? Has there been any serious difficulty in dealing with the reactance of these resistors?

In measuring the constants of the circuit, the unknown impedance is determined from a ratio of two voltages. I know from experience, however, that this method is by no means always justifiable. Can the authors give any hint as to how discrepancies may be avoided?

Finally, various parts of the authors' apparatus are screened by copper sheet, which may not be an equipotential surface at the frequencies in question. It would be interesting to know whether this factor has caused any trouble.

Mr. R. F. O'Neill (*communicated*): The paper by Messrs. Cork and Pawsey explains the importance of minimizing reflections due to irregularities in the feeder system of a high-definition television system. Since in visual reception the effects of reflection are in general more noticeable than in the case of aural reception, the requirements, as regards uniformity and correct matching, imposed on any cable link in such a system are more stringent than in the case of sound transmission.

The problem of overcoming distortion from this cause has also received considerable attention abroad, notably in Germany, where television signals are relayed by cables. The work of German investigators has been chiefly concerned with the question of random irregularities in the cable and in establishing permissible mechanical tolerances in manufacture.

Picture distortion due to echoes was also encountered in the development of facsimile equipment some 10–12 years ago, in the course of transmissions over long trunk routes. This distortion, which came to be known as "plastic," was accounted for partly by reflection and partly by phase distortion, due to non-linear phase/frequency characteristics; and to reduce these effects specially corrected cables are now employed for commercial picture transmission. An excellent treatment of the subject of "plastic" is given by Dr. Schröter in his book on picture telegraphy.*

Irregularities in cables have of course engaged the attention of telephone engineers for many years. In telephone transmission they impose a limitation on the accuracy of balance which can be maintained between the cable and the balancing network in two-wire repeater working. In this connection, attention may be drawn to a paper by Dr. Rosen,† and it is interesting to note that his analysis of the effect of one or more discontinuities is identical with that of the present paper. Among the deductions made at the end of Dr. Rosen's paper is that the extent to which neighbouring deviations in opposite directions cancel each other depends upon α, and, other things being equal, the greater the phase constant the more will be the resultant irregularity at the end.

As will be realized from the paper, much of the preliminary investigation, and its consummation at Alexandra Palace, was in the nature of field work, and the good agreement shown between theory and observation is ample testimony to the accuracy of the authors' measuring gear.

Mr. K. S. Phillips (*communicated*): Messrs. McGee and Lubszynski describe how, in the case of the Super-

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* F. Schröter: "Handbook of Picture Telegraphy and Television" (Julius Springer, Berlin, 1932).

† *Journal I.E.E.*, 1927, vol. 65, p. 989.

Emitron, the electron gun is not situated below the window of the camera, but is rotated about the axis of the window neck, to compensate for the rotation of the image due to the magnetic lens. Mr. Espley suggested at the discussion that it would lead to simpler construction if the scanning system were rotated about the electron gun, and the gun itself situated below the window, as in the ordinary Emitron. While this would mean that a complicated scan would be required, it must be remembered that, except when the gun is situated behind the mosaic, a trapezium scan must always be employed. If the retention of the electron gun exactly below the window does in fact represent an advantage, there appears to be a better method of achieving this.

Consider an electron lens in which the magnetic field along the axis at x is Hx, and the accelerating potential is V. The focal length of the lens is given by

$$\frac{1}{f} = \frac{0\cdot022}{V}\int_{-\infty}^{\infty} H_x^2 dx$$

The angle ψ, through which the resulting image is rotated, is given by

$$\psi = -\frac{0\cdot15}{V^{\frac{1}{2}}}\int_{-\infty}^{\infty} H_x dx$$

Now the sense of H_x depends on the direction of the current in the focus coil. If, then, we employ, instead of one focus coil, two adjacent coils in which the focus current flows in opposite directions, the focusing effects of the two coils will add up, while rotational effects will subtract. Hence we can arrange that the total rotation is nil. This idea was suggested by Stabenow.*

Mr. N. M. Rust (*communicated*): Messrs. Cork and Pawsey quite correctly state in the Introduction to their paper that the conditions necessary to ensure the distortionless transmission of a television signal through a feeder system have received little attention. They do not bring out, however, that the effects involved were previously well understood in relation both to telephone and to facsimile transmission along wires, and to some extent in relation to high-frequency feeder applications for commercial communications. The measurement technique employed was also a special adaptation of the well-known "impedance meter" technique used in the adjustment of commercial short-wave transmitting and receiving systems, which had the advantage of simplicity and robustness, and gave sufficiently accurate results to fulfil the requirements involved.

Comparing short-wave communication problems with television, in commercial short-wave applications the highest carrier frequency used is approximately half the television carrier frequency, whilst the sideband spectrum required even for telephony is only ± 10 kc./sec., in contrast with the ± 2·5 Mc./sec. required for television. In consequence, effects which may be of importance for television are of no practical importance for commercial communications. In quite early work with Franklin feeders it was found that the half-wave spacing of insulators must be avoided as giving undesirable effects. The quarter-wave cancellation effect was also appreciated, although it was never necessary to make practical use of it.

In some cases it was found to be desirable to correct for expansion-box irregularities. This was carried out by the very simple and practical expedient, from the point of view of the installation engineer, of using a larger-diameter inner conductor in the expansion box to bring its characteristic impedance (treated as a short length of line) down to that of the feeders which it connected together.

In relation to the brief description of the aerial system, although it is appreciated that this was from the authors' point of view regarded as incidental to the method of transformation and correction employed in matching it to the feeders, it would certainly have proved fitting to have mentioned Mr. C. S. Franklin as being responsible for its design and development.

I should like to say that the authors displayed great patience and courage when carrying out these measurements. Even in comparatively calm weather there was quite perceptible sway in the experimental mast at Hayes and in that at Alexandra Palace. In the latter place the aerial measurements were carried on considerably more than 200 ft. up in the air, and in some cases the support for both apparatus and personnel was rather precarious. Difficulties were added in bad weather owing to the slippery foothold on the metal ladders and platforms. It will be realized to some extent what conditions were like if it is mentioned that safety belts had frequently to be used.

THE AUTHORS' REPLIES TO THE DISCUSSION

Messrs. E. C. Cork and **J. L. Pawsey** (*in reply*): Mr. Bedford has raised the fundamental point as to the degree to which the feeder and aerial must be free from reflection phenomena. The problem arose from the observation of striae in pictures due to reflections occurring in what would be considered to be a reasonably broad-band aerial system. The aim of this work was to reduce this form of distortion below the limits of visibility, allowing for future improvements in definition of receivers up to the useful limit of the system. Estimates along the lines of the paper indicate that the permissible degree of mismatch is small. Observations on high-definition receivers using long low-loss aerial feeders have frequently shown this form of distortion, which disappears when a broad-band matched-aerial feeder system is used. With regard to the possibility that the accuracy of the results given may be of an artificial nature, the fluctuation of resistance with frequency constitutes a fairly direct measure of the reflected waves arriving at the measuring point. The absolute measure of the input impedance is relatively less accurate than the fluctuation, but the final results do not depend on any absolute value of impedance. Indeed, a completely satisfactory definition of the impedance between two points separated by a finite distance appears to be lacking.

We are interested to learn from Mr. Birkinshaw that

* *Zeitschrift für Physik*, 1935, vol. 96, p. 634.

the insulation resistance of the feeder maintains a high value. Great care was taken as to the scrupulous cleanliness of the insulators and the sealing of the angle boxes. Without such precautions large fluctuations can take place. In the course of the erection of the feeder various slight mismatches were taken up by the methods indicated in the paper.

In reply to Mr. Espley, the feeder selected was one of the standard types available, and has an ample margin of safety. There is no significance in the radius ratio 3·6 in this case. The effect of the proximity of the sound aerial was measured, the voltage induced into the vision aerial being about 1 % of that in the sound aerial. With regard to series resistance developing in expansion joints, we were apprehensive of this trouble, but with the type employed no faults due to this cause were encountered. We have no information as to the steady-state propagation characteristics of the transmitter.

With regard to the reduction of the effects of eccentricity discussed by Mr. Mumford, the necessary insulator clearance and consequently the maximum eccentricity is determined by the possible ellipticity and lack of uniformity of the pipes, so as to ensure that the insulators are always free. Probably in a future installation sufficiently uniform pipes will be obtainable to allow considerably smaller clearances. Small correcting condensers were occasionally inserted in the Alexandra Palace feeder to correct for eccentricity.

With reference to the degree of maintenance of the characteristic curves of the feeder and aerial, a check measurement made about a month later showed no appreciable change. Further, since observations of the transmitted picture have shown no evidence of distortion due to this cause, it is probable that the system has remained substantially unchanged. As Mr. Mumford has pointed out, the aerial produced a mismatch much greater than that due to irregularities in the feeder. In the initial experiments a centre-fed dipole aerial of about 2 ft. diameter was developed and erected about 15 ft. above the top platform of the Hayes mast. The resistance variation over 4 Mc./sec. was 24 %, and the reactance variation 20 % of the tuned resistance value. When this aerial had been corrected by a suitable tuned circuit the maximum impedance deviation from the characteristic impedance was $\pm 2 \cdot 5$ %.

We agree with Dr. Jarvis as to the fundamental nature in television of the time response of circuits as opposed to frequency response. From this point of view the analysis of a circuit may both practically and theoretically be facilitated by the use of pulse technique. The practical application of such technique requires more elaborate apparatus than the simple measuring circuit indicated in the paper. With regard to the variation of the output of the transmitter, measurements of the overall response have been made over the range of modulation frequencies, using a receiver at a distant point. The variation of the output is complicated by the existence of the two sidebands, which have different amplitudes due to the non-symmetrical form of the overall aerial-feeder characteristic. The measurement is the resultant of a large number of factors, with the result that the overall accuracy is not sufficient to show a variation of a few per cent due to the aerial.

Mr. Sandeman has raised the question of the temperature control of the measuring resistances. These were frequently measured by means of a bridge on the spot and showed little variation. The measuring circuit was adopted after considerable testing and correction for the residual errors of leads. Bridge methods were tried, with the consequent increase of reading accuracy owing to the use of a null detector, but this accuracy can be fictitious and we have not so far replaced the substitution method for general use.

We agree with Mr. Ward that it will be necessary to lay down specifications and methods of testing for high-frequency cables, which are already numerous. Provisionally we specify the cable in terms of its characteristic impedance, attenuation, and velocity constants. It seems desirable to have some figure describing its uniformity and also a comparison with a cable with copper conductors and a loss-free dielectric, probably air. We have no information of the magnitude of the increase of attenuation due to copper oxide. It appears to be a fortunate accident that resistances of the type used were approximately independent of frequency up to at least 50 Mc./sec. Further, they are substantially resistive, the change-over between the capacitive reactance of very high resistances and the inductive reactance of very low ones occurring somewhere in this region. Errors may arise from several causes in deducing resistance from a ratio of voltages. Those due to the unknown law of the voltmeter and the change of loading on the oscillator are evaded by the substitution of an equivalent resistance. Errors arising from the reactance of leads must be reduced to small magnitude in the design of the apparatus. Considerable errors may arise from potential differences over the surfaces of metallic sheets, on account of the finite impedance of the surface. These potentials may be reduced by adequate screening and the design of the apparatus to avoid induced e.m.f's.

We thank Mr. O'Neill for pointing out Dr. Rosen's treatment of irregularities in cables, which is a fuller treatment of Section (3) of the present paper.

In reply to Mr. Rust's remarks, of course reflection problems in long telephone lines have been dealt with for many years, and cases have arisen—particularly in long-wave aerials, where the band-width is a reasonable percentage of the carrier frequency—requiring very careful building-out of the aerial impedance; but as far as we know this is the first occasion on which the double problem has been dealt with, where not only is the band-width a sufficient percentage of the characteristic frequency to require careful treatment of the aerial, but also the aerial feeder is long with reference to the modulation band-width. With regard to the aerial, we have stated that it was designed by the engineers of the Marconi Co. and are glad to join with Mr. Rust in acknowledging the part played by Mr. Franklin in this work.

Messrs. J. D. McGee and **H. G. Lubszynski** (*in reply*): Replying to Mr. Bedford, the transmission of film is more difficult than the transmission of studio scenes, because of the frequent rapid changes in illumination encountered with the former. These changes require frequent and rapid readjustment of "tilt" correction. The intermittent scanning method for film transmission was not discarded owing to an insufficient "memory" of the Emitron, but because with this type of scanning large additional undesired pulses appear between frames and complicate the somewhat difficult conditions still further. From this point of view the continuous-motion projector offers more advantages than disadvantages.

In reply to Mr. Birkinshaw, the type of mosaic with low red sensitivity now in use in the Emitron differs considerably from the old type of mosaic described in the paper. This new mosaic will be described in a future publication by Dr. L. Klatzow.

Replying to Mr. Espley, the electron-lens correction for keeping the scanning beam in focus over the whole of the mosaic area was never used in practice because it was found possible to obtain sufficient depth of focus of the electron beam to cover the whole mosaic by stopping down the apertures which limit the electron beam. The method could, however, be used in both frame and line direction.

The compensation of the image rotation by rotating the deflecting fields would complicate the scanning circuits. Owing to the oblique incidence of the scanning beam on the mosaic, the scanned patch shows a keystone distortion when the line-scanning amplitude remains constant over the whole frame period. This distortion is compensated electrically by increasing the amplitude of the line-scanning currents from the top to the bottom of the mosaic. If the deflection fields were rotated a second compensation would have to be introduced in the frame-scanning circuits, which would complicate matters considerably.

The spot of low sensitivity which sometimes appears in the centre of the picture is due to burning of the sensitive surface by the electron beam, which has been allowed to remain stationary for some time. Safety devices are now used which cut off the electron beam if the scanning fails.

No difficulty is experienced due to "poisoning" of the mosaic by negative ions if satisfactory vacuum conditions are maintained.

In reply to Mr. Mumford, the noise in the picture from an Emitron or Super-Emitron is composed of three components: (1) the noise of the first stage of the amplifier; (2) the noise carried by the scanning beam; and (3) the noise of the photo current. Of these, the first is the determining factor. The noise contributed by the scanning beam is, at present, about one order of magnitude below the amplifier noise. The signal output must be so large that the signal is not drowned by the noise of the amplifier. Under these conditions the noise of the photo-current is negligibly small. When reference was made to the sensitivities of the two types of tube, their signal output in volts at equal light flux into the tubes was meant. Therefore, the statement that one type of tube is n times more sensitive than the other means that its signal/noise ratio in volts is n times better for equal light flux. As in these tubes the signal output is proportional to the light input only up to a certain intensity of illumination, and as the signal reaches a saturation value for strong illuminations, it is more correct to state that if one type of tube is n times more sensitive than another their signal/noise ratios in volts are equal when the light input into one is n times smaller than that into the other.

The signal output can be increased by increasing the beam current, and consequently the signal/noise ratio could be improved considerably before the noise due to the scanning beam became serious, but with increasing beam current the "tilt" signal becomes unmanageable without elaborate correcting circuits. Thus the spurious "tilt" signal ultimately sets the limit to the signal/noise ratio obtainable.

The inherent "tilt" is approximately the same for the same beam current in both types of tube, and hence their signal/"tilt" ratios are to one another as their sensitivities. The greater signal/noise ratio of the Super-Emitron is mainly due to the more efficient conversion of the light image into an electrostatic charge image on the mosaic.

The Super-Emitron is not used universally by the B.B.C., for several reasons. The few cameras and tubes at present available are largely experimental and it was considered inadvisable to make a large number of either until they had given convincing proof of their superiority over standard Emitrons, of which a large stock must be kept on hand. Consequently the Super-Emitron cameras have been used mainly for outside broadcasts which would have been difficult, if not impossible, for the standard Emitron. In these cases the geometrical distortion of the picture transmitted by the Super-Emitron was tolerated in order to get a picture at all, while the standard Emitron with its more accurate geometry of picture is preferred for studio work, where adequate illumination can be provided. Improvements in design which are being introduced into new Super-Emitrons will, we hope, enable them to displace the old cameras from the studio, with considerable improvement in the pictures.

Dr. Gabor's objection to accepting our interpretation of Campbell Swinton's suggestion appears to be based on the erroneous impression that Campbell Swinton had in mind only photo-conductive mosaics. That this is not so is clear from his address to the Röntgen Society,* in which he describes a mosaic as follows: "The metallic cubes which compose J (the mosaic) are made of some metal such as rubidium. . . ." This would be a photo-emissive mosaic which would have no time-lag and would be capable only of charge storage.

Pincushion distortion is not a defect of electron images formed by electromagnetic electron lenses. Hence no correction is necessary.

With regard to the secondary emission from carbon, the maximum secondary emission from pure carbon is very low indeed. It was found, however, that carbon adsorbed caesium very readily, and then secondary coefficients as high as 9 secondaries per primary were obtained.

In reply to Mr. Bridgewater, all the types of distortion described are particular forms of "tilt." These spurious signals depend on several factors, such as the distribution of illumination on the mosaic, the nature of the mosaic surface, small stray magnetic fields, and so on.

The type of distortion which shows as a white edging extending into large areas of black in the picture is due to the "tilt" signal assuming a wave-form which cannot easily be corrected by the simple correcting circuits that are used. The image can be improved by increasing the illumination, since then less amplification is required and a better signal/tilt ratio is obtained.

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* *Journal of the Röntgen Society*, 1912, vol. 8, p. 11.

The dark band which sometimes appears at the top of the picture is due to an accentuation of the strong black pulse which occurs at the beginning of each frame. When the scanning beam begins a new frame, it is surrounded by a region of the mosaic which is at a more negative potential than normal. Hence the secondary electrons liberated by the beam tend to return to the collecting electrode in greater numbers than usual. Thus, although picture signals are produced the black level is depressed, and it is only after the spot has scanned a number of lines and so built up a normal distribution of potential around it that the black level of the picture reaches its normal value. This edge effect is usually suppressed in the amplifiers, but occasionally under unfavourable conditions it may extend into the picture. To reduce this effect to the minimum the scanning beam should be allowed to start scanning the mosaic immediately it has completed its return stroke, so that it may have as much time as possible to build up the normal charge distribution on the mosaic before the amplifier suppression is removed.

The type of distortion known as " streaking " may be explained as follows. The production of undistorted picture signals depends on the assumption that the peak positive potential, to which the elements of the mosaic are brought by the scanning beam, is constant and the same at all points of the mosaic, and that a constant fraction of the secondary electrons liberated from a point return to the second anode. That this is not strictly true is shown by the appearance of spurious signals such as " tilt." It is clear that the peak positive potential to which the mosaic elements are charged cannot vary rapidly, otherwise fine detail of gradation in the image would not be reproduced. Hence it should follow that when the electron beam scans a positively charged (white) part of the mosaic, the peak potential will rise a little higher than when it scans an uncharged area. While such areas alternate rapidly, they will have very little effect on the peak potential of the scanned spot. If, however, the spot remains on such a charged area (for example, a white band running parallel to the scanning lines across the picture) for a considerable part of one line, the peak potential will drift more positive and the amplitude of the signal will decrease. When the scanning spot passes off the white band on to the black surrounding area it will initially produce a signal of too great an amplitude in the black direction which fades out in approximately the time taken to scan one line.

The comparatively slow rate at which this peak potential drifts under the influence of charged areas on the mosaic is probably due to the control exerted on the potential to which each mosaic element will rise by the potential of the element that has just been scanned; that is, the most positive area in the neighbourhood. This tends to keep the peak positive potential constant, and it is only slowly influenced by charged areas on the mosaic.

In reply to Mr. Sandeman, the maxima of the spectral response curves of the Emitron and Super-Emitron are in the yellow-green, and the whole response curves are very close to that of the human eye. There is very little sensitivity in the infra-red. Both types of tube can, however, easily be made so that they have their maximum sensitivity in the red, and their response extending in the infra-red to about 12 000 Å.U.

Replying to Mr. Phillips, the two-coil method of obtaining images without rotation, suggested by Stabenow,* was tried but was found unsatisfactory. Though an upright picture was obtained over a small central zone, the spiral distortion at some distance from the axis was so enormous that the rest of the image was spiralled into a bright ring surrounding the central area.

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* *Zeitschrift für Physik*, 1935, vol. 96, p. 634.

THE DESIGN AND CONSTRUCTION OF A SHORT-WAVE FIELD-STRENGTH MEASURING SET*

By F. M. COLEBROOK, B.Sc., and A. C. GORDON-SMITH.

(*Paper first received 22nd August, and in final form 19th October,* 1938.)

SUMMARY

The paper contains a discussion of the principal features of the design of apparatus for the measurement of field strength at very short wavelengths by means of a loop aerial. The apparatus consists of a high-gain intermediate-frequency amplifier (about 1 Mc./sec.) associated with a signal-frequency and frequency-change unit of constant conversion efficiency. The e.m.f. of thermal agitation in the input tuned circuit of the intermediate-frequency amplifier is used as a standard signal for setting the amplifier to a known gain. The combination of intermediate-frequency amplifier and signal-frequency unit is calibrated over its signal-frequency range by means of a radiator which gives a horizontally-polarized radiation of calculable intensity.

It is shown that a single-phase, i.e. unsymmetrical, frequency-change system has certain practical advantages over biphase systems, and that this can be used without loss of the necessary balance of the aerial system, if the latter is coupled inductively through a Faraday screen to the frequency-change circuit. The detailed design of the various couplings and of the diode frequency-conversion circuit are analysed in Appendices.

The intermediate-frequency amplifier embodies continuously-variable gain-controls, calibrated in decibels, these components being a particular type of commercial carbon-track " volume control " which has been found satisfactory for this quantitative use.

The set can be used for the measurement of field strengths down to the order of microvolts per metre at wavelengths from 7-11 metres (27-43 Mc./sec.) with the frequency-change and aerial units described.

(1) THE TECHNIQUE OF SHORT-WAVE FIELD-STRENGTH MEASUREMENT

There are at present two principal methods of measuring an unknown field-intensity: (*a*) the measurement of the voltage produced by the unknown field in some suitable form of aerial, and the subsequent calculation of the corresponding field-intensity in terms of the constants of the aerial; (*b*) the comparison of the unknown field-intensity with a known field-intensity by means of some form of transfer instrument—in practice, an aerial and a receiving set.

The first method involves a radio-frequency voltmeter of very high sensitivity, i.e. capable of the measurement of potential differences of the order of microvolts. In practice this means an amplifying radio-frequency receiver of very high gain, the amplitude calibration of which is determinable by the application of a known small voltage of signal-frequency to its input terminals, or in series with the aerial connected to its input terminals. Until a few years ago, some form of this method was the only one employed in practice, but its reliable application was limited to measurements of fields with wavelengths greater than about 15 metres, owing to the technical difficulties of the calibration process referred to at wavelengths shorter than this. Recently, however, there has been produced by a well-known firm an equipment in which the range of application of this method has been extended to about 3 metres.

A theoretically possible variation of the first method is a combination of a high-gain intermediate-frequency amplifier, the performance of which is readily determinable by known methods, with a signal-frequency collector and frequency-change unit of directly calculable performance. The authors attempted a design based on this scheme, chiefly in order to secure an independent check of the second or known-field method. The attempt failed, for reasons which are themselves of sufficient interest to justify the following brief account.

The signal-frequency unit consisted of a symmetrical loop aerial, directly coupled to a frequency-change unit, which, to preserve symmetry and to eliminate " antenna effect in the loop aerial, must be of the biphase or push-pull type. The effective height of the loop aerial (i.e. $2\pi \times$ area turns divided by wavelength) was calculated, and its Q or resonance-factor ($\omega L/R$) determined by a reactance-variation method. Considering now the frequency-change part of the unit, it is shown in Appendix III that the quantitative relationships of frequency conversion by means of a diode circuit are theoretically very simple and are amenable to direct measurement. The circuit finally employed is illustrated in Fig. 1, but in a first construction the oscillator voltage was mistakenly introduced into the common lead in series with the tuned intermediate-frequency circuit. An examination of the theory of such a circuit shows that, assuming perfect symmetry, the output intermediate-frequency voltage should have been zero. In practice the output obtained was so large that the error in the circuit was overlooked for some while. This led to a more systematic examination of the extent to which the postulated symmetry could in fact be realized at very short wavelengths. A unit was constructed in which two intermediate-frequency circuits were incorporated, with zero mutual inductance between them. The first was arranged as shown in Fig. 1, and the second in the position shown as the local-oscillator coupling in Fig. 1, the local-oscillator coupling being transferred to the common lead. With perfect symmetry the whole of the intermediate-frequency output would appear across the second circuit, none appearing in the first. It was found that although all normal steps were taken to ensure electrical symmetry, the output in the wrongly connected circuit was nearly twice as great as that in the correctly connected circuit. Even with an auxiliary 3-plate balancing condenser of small capacitance, introduced as shown in Fig. 1, the necessary complete symmetry was not obtained, for the output in the wrongly connected circuit was not reducible below about half of that given by the other. It was thus apparent that very elaborate precautions would be required, combined with some degree of auxiliary balancing by additional variable impedances, in order to realize to an adequate extent the electrical symmetry postulated in the simple theory. In view of these experiments it seemed doubtful whether a directly calculable equipment would in fact be realized with sufficient certainty to justify its use as an independent check of the known-field method. The attempt was therefore abandoned, and this brief account is only included for information in relation to any future work on these lines, and as a warning against the too-facile assumption of actual electrical symmetry in apparently symmetrical dispositions at very short wavelengths.

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* Official communication from the National Physical Laboratory. Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 388.

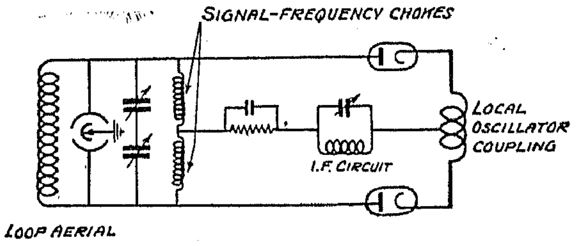

Fig. 1.—Biphase or "push-pull" diode frequency-change circuit.

The second method depends essentially on the calculation of the field intensity produced by some simple form of radiator in terms of the physical constants of the radiator (in practice, a closed-loop aerial), the current in the radiator, the electrical constants of the ground, and the disposition of the point of measurement and the radiator relative to each other and to earth. The method was originally introduced when measurements on the shorter wavelengths became of commercial importance, as a way out of the difficulty already mentioned in connection with the first method. In its earliest form, e.g. as described by Schelling, Burrows, and Ferrell,* the technique involved some elements of uncertainty owing to the difficulty of making accurate allowance for the effect of the earth. Recently, however, an important advance has been made in this field by J. S. McPetrie and B. G. Pressey, who have called attention to the valuable simplifications which are introduced into the underlying theory by the use of horizontally polarized waves instead of the vertically polarized waves originally employed. Indeed, with this simplifying modification it may be argued that the standard-field method, at least in the range from, say, 15 metres downwards, is as free from technical and theoretical uncertainties as the alternative, or standard-microvolt, method, and can therefore claim at least as high a status in practice even though in theory it may be one stage further removed from the fundamental standards.

The apparatus described in the present paper is intended for use in conjunction with a standard-field radiator as described by McPetrie and Pressey.* The intention of the design is, however, that a single calibration of this kind over its working range (7 to 11 metres) should suffice. That is to say, the high-gain intermediate-frequency part of the equipment is adjustable to a standard condition by reference to the e.m.f. of thermal agitation in the input intermediate-frequency tuned circuit, and the signal-frequency part is designed for inherent constancy within practical limits.

It is desired to emphasize that the apparatus is described primarily for the information of those working in this field, and not as a standard and perfected equipment, and the description is therefore more concerned with the principles involved than with the detail of this particular assembly.

(2) GENERAL DESCRIPTION

The equipment consists of a unit comprising a tunable frame aerial, a frequency-change oscillator, and a diode frequency-conversion circuit. This unit is coupled by means of a concentric screened lead to a second unit consisting of a high-gain intermediate-frequency amplifier and rectifier. The tuning range of the signal-frequency unit is from 7 to 11 metres (43 to 29 Mc./sec.), and the intermediate-frequency amplifier is tuned to 250 metres (1·2 Mc./sec.). The necessary voltage supplies are derived from a dry battery (120 volts) and accumulator (4 volts) contained in a separate wooden box, with plug-in connectors to the two units. The normal disposition of these units is shown in the photograph (Fig. 2). The units are described in detail below.

(3) THE SIGNAL-FREQUENCY UNIT

(a) Aerial

In a previous paper,* one of the authors has analysed the behaviour of single-turn aerials in relation to direction-finding and field-strength measurement on short wavelengths. For present purposes, the most important conclusions are:—

(i) If the lowest point of such a loop aerial is connected to earth or to a point having appreciable capacitance to earth there will be, in general, even if the coil and tuning system are electrically and geometrically symmetrical, appreciable current flowing in such connection and it is desirable that this shall not affect the terminal potential-difference of the aerial. Thus, if the aerial is tuned by a three-terminal symmetrical condenser at the bottom of the loop, it is desirable that the receiver shall record the vector sum of the potential differences across each half of the tuning condenser. In a symmetrical system, this vector sum will be independent of any current flowing in the earth or "earthy-point" connection. Alternatively, and preferably, the aerial can be so connected as to have a maximum of impedance to earth at its electrical centre.

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* See Bibliography, (1).

* See Bibliography, (2). † *Ibid.*, (3).

(ii) It is not essential that the periphery of the loop shall be small compared with the wavelength of operation, i.e. it is not essential that the current in the loop shall be uniform throughout the length of the loop. It appears that for peripheral lengths less than about a quarter of the wavelength of operation, neither the induced e.m.f. nor the polar-distribution of " effective height " is very sensitive to current distribution. If necessary, the effect of non-uniform current-distribution on the induced e.m.f. can be calculated and allowed for, but if, as in the present case, the set is calibrated by a standard field radiator, the effect of current distribution on effective height does not enter into the matter. If a substantially uniform vertical polar-distribution of reception-sensitivity is desired, the peripheral length should not exceed about one-fifth of the wavelength of operation.

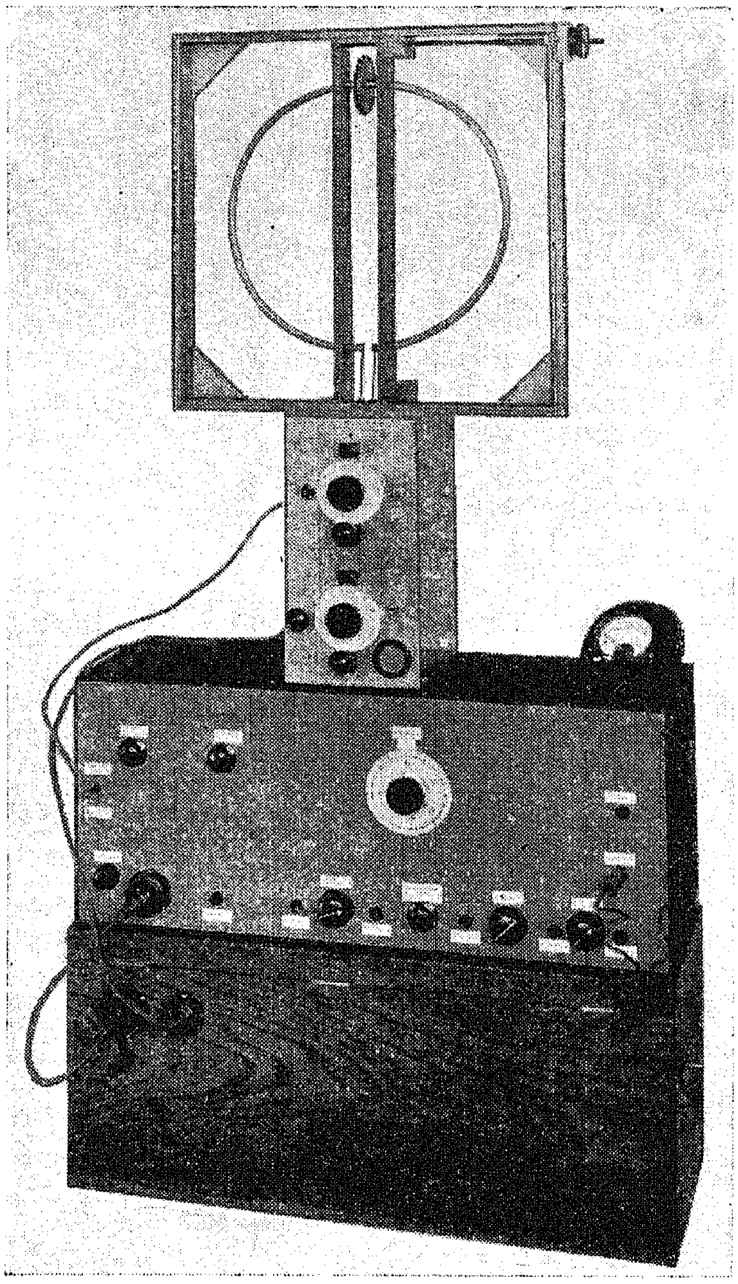

Fig. 2

A practical point of some importance is that the aerial shall have a minimum of " pick-up " with respect to waves of frequencies in the region of that of the intermediate-frequency amplifier. For this reason the coupling of the aerial to the receiver should be as loose as possible consistent with signal-frequency efficiency.

A means of satisfying these various requirements is illustrated schematically in Fig. 3. The aerial is tuned by a variable capacitance at the top of the loop (a system adopted in the commercially-produced equipment already referred to) and is inductively coupled to a closed secondary circuit, with a metal screen between the windings. A convenient screened coupling was obtained by metallizing the external cylindrical surface of a ribbed ceramic coil-former (the ribs being left unmetallized); the secondary was wound externally on the ribs of the former, and the small single-turn coupling coil was inserted inside. The coil former was mounted so that its upper open end was against the roof of the metal screening case, as indicated in Fig. 3. (The metal case in this model was constructed of tinned iron of fairly heavy gauge. In a later model, copper plate is being used.)

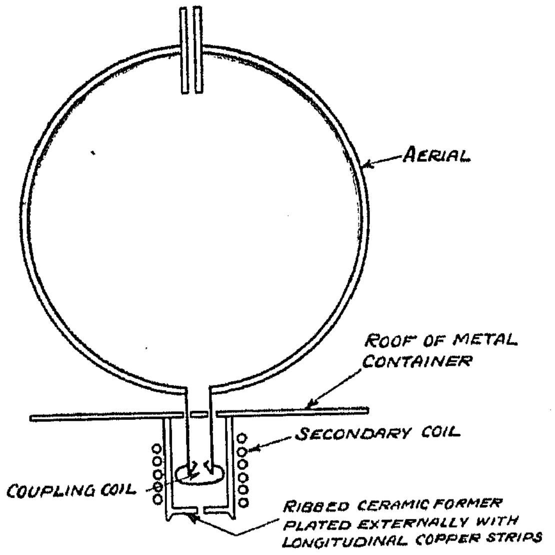
AERIAL
ROOF OF METAL CONTAINER
SECONDARY COIL
COUPLING COIL
RIBBED CERAMIC FORMER PLATED EXTERNALLY WITH LONGITUDINAL COPPER STRIPS

Fig. 3.—Assembly of loop aerial, screened coupling, and secondary coil.

It is shown in Appendix I that the optimum value of the coefficient of coupling is given by

$$k = \sqrt{(p_1 p_2)}$$

where p_1 and p_2 are the " power factors " of the aerial and secondary circuit respectively. The optimum is not critical, and can be approximated to by trial. It cannot in any case be realized at all frequencies in the range, and the condition to be aimed at is that the coupling shall not be appreciably below the best value at any point.

An important advantage of the screened coupling system described is that it avoids the necessity for biphase or " push-pull " frequency-conversion. That is to say, it enables a single-phase or unsymmetrical frequency-conversion circuit to be used without impairment of the essential symmetry of the loop-aerial system. The

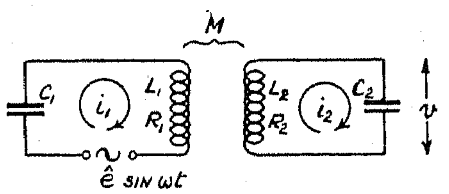

Fig. 4

difficulties associated with symmetrical valve circuits at very short wavelengths have been referred to in Section (1).

It must be noted that the tuning process with these loosely coupled tuned circuits is not, unfortunately, a simple matter of adjustment of the two tuning condensers in turn. Referring to Fig. 4, for any setting of C_1 there will be a tuning value for C_2, but C_1 must then be displaced one way or the other and C_2 retuned, this being repeated until the absolute optimum is reached.

The type of aerial loop used in this and other models has proved very satisfactory. It will tune over an unusually wide range of wavelengths (at least $3{\cdot}5:1$), minimizes dielectric loss, and is comparatively simple in construction. In the present model it consists of two semi-circles of copper tube, mounted so that one of the semi-circles hinges at the bottom. The tuning condenser consists of two brass discs soldered directly on to the top ends of the two semi-circles. The end of a threaded ebonite rod presses against the plate soldered to the pivoted semi-circle, and tends to close the gap against the tension of a rubber band.

(b) Frequency Conversion

It was decided to use a diode frequency conversion circuit, for the following reasons:—

(i) Linearity.

With a sufficiently large voltage from the frequency-change oscillator, diode frequency conversion is linear with respect to signal voltage, and the intermediate-frequency output is substantially independent of oscillator voltage. This is a valuable feature from the point of view of maintenance of calibration.

(ii) Quantitative relationship.

It is shown in Appendix III that the quantitative relationships involved are comparatively simple.

(iii) Harmonic operation.

It has been shown by Strutt* that in diode circuit frequency-conversion with a large oscillator voltage (5 volts or more), the conversion efficiency using the second or third or even higher-order harmonics of the oscillator output in the diode circuit should not be materially less than for the fundamental. The use of harmonics in this way has advantages in respect of "pulling" and interaction of controls. In the present model the fundamental is actually used for the normal range, but it was found possible to extend the range downwards to about 4 metres by short-circuiting part of the closed secondary circuit and using the second harmonic of the oscillator output.†

The circuit finally adopted, after considerable experiment, is shown in all essentials in Fig. 5. The coupling condenser C_1 is intended to lessen the load of the diode on the closed secondary circuit by a kind of matching process. It can be shown that if the effective resistance of the diode is R_d and the dynamic resistance of the tuned circuit is R_c, the optimum value of C_1 is

$$\frac{1}{\omega\sqrt{[R_d(R_c - R_d)]}}$$

At very high frequencies this is, in general, a very small quantity, e.g. with $R_d = 10^4$ and $R_c = 2 \times 10^4$, $C_1 = 0{\cdot}4\ \mu\mu\text{F}$ (approximately) at 7 metres. The best value will in practice exceed this calculated value on account of the anode-to-cathode capacitance of the diode, but will generally be less than 1 $\mu\mu$F. For the present model a length of "styro flex" tubing was slipped over a short piece of wire, which was then inserted in a brass tube of about 2 mm. bore, making a tiny variable concentric-cylinder form of condenser.

Condenser C_3 is similarly a matching condenser for the intermediate-frequency load circuit. The frequency being lower (about $1{\cdot}2$ Mc./sec.), a somewhat larger value is appropriate.

The frequency-change oscillator is coupled to the diode by means of the small capacitance C_2. In practice it was found that sufficient coupling for an induced diode-circuit voltage of 5 or 6 volts was obtained by mounting a short straight wire along the length of the oscillator coil and about $\frac{1}{4}$ in. away from it. Various other forms of coupling were tried, including "electron" coupling, using an "acorn" pentode valve as the oscillator and coupling valves, but the arrangement shown in Fig. 5 was found to be the most satisfactory in respect of both "pulling" and available amplitude. The smallness of the capacitance C_1 contributes materially to the success of this mode of coupling.

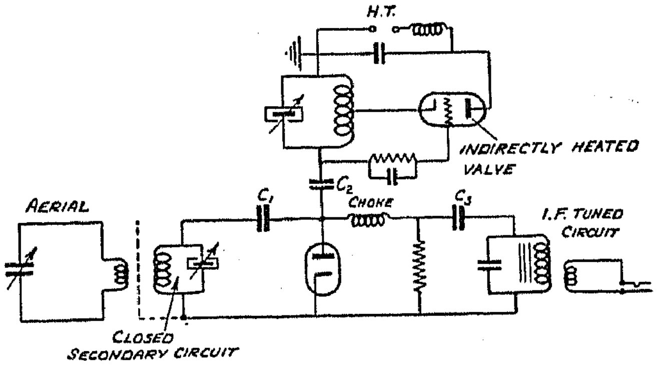

Fig. 5.—Circuit diagram of aerial and frequency-change unit.

The oscillator circuit used is a somewhat unfamiliar but very useful variant of the Hartley circuit, made possible by the use of an indirectly heated valve. The oscillatory circuit is essentially in series with the supply voltage and the valve, and chokes are therefore not essential for oscillation. Further, one electrode of the tuning condenser can be earthed. In practice, a choke can be inserted in the H.T. lead, but this is merely to restrict the flow of radio-frequency current in the supply leads.

The intermediate-frequency coil is wound on a small iron-dust core, and is provided with a secondary winding of a few turns for matching the tuned impedance to the surge impedance of a concentric screened line for coupling to a similar tuned circuit in the intermediate-frequency amplifier. In general, a short screened coupling lead is used and the signal-frequency unit stands on the intermediate-frequency amplifier, as shown in Fig. 2. If desired, however, a longer coupling lead can be used and the signal-frequency unit can be remote from the intermediate-frequency amplifier. It was found that the matching was sufficiently good to give substantially the same output with the long as with the short coupling lead.

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* See Bibliography, (4).

† Since the present paper was written, further field-strength measuring equipments for use at wavelengths down to 1 metre have been developed by the authors. The circuits employed are essentially the same as those described in the present paper, with tuned dipoles and "flex" transmission-line couplings in place of the tuned loop aerial. Diode-circuit frequency-conversion in the harmonic mode of operation has been used throughout and has proved satisfactory. The calibration at 3 metres was checked after some months of field use and was found not to have changed materially.

An "acorn" triode valve is used for the oscillator circuit, and a similar valve, with anode and grid connected, is used as the diode. (The set was constructed prior to the recent introduction of certain very small diode valves, now commercially obtainable, which would otherwise have been used.)

(4) THE INTERMEDIATE-FREQUENCY UNIT

The circuit diagram of the intermediate-frequency amplifier is shown in Fig. 6. It comprises four stages of amplification by means of radio-frequency pentode valves, a diode rectifier with an auxiliary diode for optional automatic gain-control, and a balanced 2-valve bridge-type output circuit. The tuned frequency is about 1·2 Mc./sec.

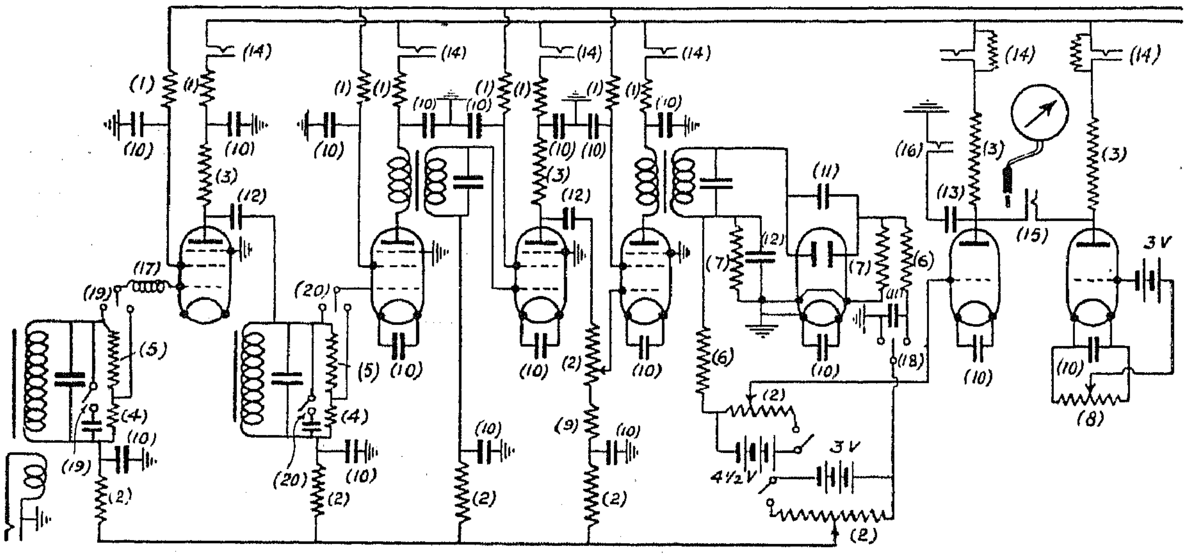

Fig. 6.—Circuit diagram of intermediate-frequency units.

(1)	5 000 ohms.
(2)	10 000 ohms.
(3)	20 000 ohms.
(4)	500 ohms.
(5)	100 000 ohms.
(6)	100 000 ohms.
(7)	1 000 000 ohms.
(8)	1 000 ohms.
(9)	500 ohms.
(10)	0·1 μF.
(11)	0·001 μF.
(12)	0·01 μF.
(13)	2 μF.
(14)	Test plug.
(15)	Output plug.
(16)	Telephone plug.
(17)	Radio-frequency choke (7–10 m.).
(18)	Switch for A.V.C.
(19, 19)	Double-pole 2-way switches.
(20, 20)	Double-pole 2-way switches.

Many features of the design and assembly are quite normal. Only special and distinctive features will be described in detail.

(a) Use of Battery Valves

Battery valves are used for the sake of weight economy in voltage supplies. Special precautions are, however, required to eliminate interaction by way of filament circuits. Each filament is short-circuited by a 0·1-μF condenser and is connected to the common 2-volt supply by a separate twisted pair. There remained, nevertheless, a slight residual instability, which, for reasons not yet understood, disappeared when the by-pass condenser was disconnected from the first valve. It cannot be stated that the means adopted are in fact the best general means for preventing filament-circuit interaction, but they proved sufficient in the present case.

(b) Buffer Stages

There is much to be said for the use of resistance-coupled buffer stages interposed between tuned-circuit stages, even with screened valves. The first and third stages are of this kind, though a relatively flat tuned circuit is used in the first in place of a grid resistance, mainly to minimize the effect of the input capacitance on the attenuator step which is located at this point.

(c) Attenuators

There are three attenuator stages—34 and 50 db. in single steps—and a continuously variable stage of 38 db. maximum. The single-step attenuations are obtained by means of resistances connected across the first- and second-valve input tuned circuits. Double-pole 2-way switches are used, one way introducing a small preset capacitance to maintain the tuned condition when the input capacitance is removed from across the tuned circuits in the "attenuator in" position.

The continuous variation of attenuation is obtained by means of a carbon-track variable resistor of 20 000 ohms (maximum). It is a commercially obtainable type in which the moving arm makes contact by means of three separately-acting metal-spring brushes. Preliminary examination had shown that the contact so obtained is sufficiently good, definite, and reproducible for quantitative use and calibration to an order of accuracy adequate for the present purpose (1 or 2 per cent). It has obvious advantages over the alternative of fixed resistors and a multi-way step switch, and has proved quite satisfactory in practice. A small fixed resistance is connected in series, to improve the law and limit the range to about 40 db.

(d) Rectifier Circuits

The rectifier is arranged in such a way as to produce a reduction in the anode current of the output valve when a signal is received. The chief practical advantage of this arrangement is that it limits the current through the output instrument, which can never exceed the initial steady anode current of the output valve. This is about 1·5 mA, so that the maximum possible overload on the 500-μA output instrument is only 3 times full scale.

It was found necessary to derive the automatic-gain-control voltage from a separate anode in the double-diode rectifier, to prevent minute grid currents in the amplifying valves from affecting the grid potential of the output valve. This separation of functions also facilitates control of the output/input law with automatic gain control.

(e) Output Circuit

A bridge circuit is used, with independent control of the grid bias of each valve. The output instrument (0–500 μA) is not permanently attached, but is connected to a plug. The same instrument is used for inspection, shunts being permanently wired across the plug points where necessary.

The preliminary adjustment of the output circuit is carried out as follows: The instrument is plugged into the anode circuit of the last (balancing) valve and the anode current adjusted to a scale reading of 500 μA (an actual value of 1·5 mA). The instrument is then plugged into the working position and, with full attenuation in, the deflection is reduced to zero by adjustment of the grid bias of the output valve. This procedure ensures constancy of initial current in the output valve.

(5) OPERATION AND CALIBRATION

(a) Adjustment of Intermediate-frequency Amplifier

The intermediate-frequency amplifier is adjusted to a standard gain by the method introduced by the Radio Branch of the Post Office.* The frequency-change unit is first disconnected, and full attenuation is switched in. The output valve circuits are adjusted and balanced as described. The attenuator switches are all set to the minimum value, and the first input tuned circuit is brought into tune with the other tuned circuits. The manual gain control is now adjusted so that the output instrument reads 400 μA, this output being due to the e.m.f. of thermal agitation in the first tuned circuit.

If the first tuned circuit is distuned, or the first attenuator step switched in, the output falls to 90 μA, and is further reduced to practically zero if the first valve is switched out or removed. Thus practically the whole of the noise output of the amplifier (with minimum attenuation) is due to the e.m.f. of thermal agitation in the first tuned circuit, the remainder being due to shot-effect noise in the first valve. This large preponderance of the first-circuit noise is due to the fact that the first circuit is of high dynamic resistance (about 125 000 ohms, with about 100 000 ohms in parallel due to the first attenuator step), and also to the low anode currents of the valves in the standard condition (about 0·2 mA per valve).

(b) Field-strength Calibration and Measurement

The technique of the production of a known field-intensity by means of a closed-coil radidator has been fully described by McPetrie and Pressey.† For present purposes it is assumed that such a known intensity ϵ_0 is available, and that when the set is tuned to it a total attenuation N_0 is required to reduce the output-instrument indication to a certain standard value K (400 μA in the present instance). From the attenuation N giving the same standard output with an unknown field intensity ϵ, the latter can be calculated in the following way.

It will be assumed in the first instance that ϵ and ϵ_0 are so large that the contribution in each case of the "noise" voltage to the standard output is negligible. Then, if a_0 and a be the amplification values corresponding to ϵ_0 and ϵ respectively,

$$\frac{\epsilon}{\epsilon_0} = \frac{a_0}{a}$$

or

$$\begin{aligned} 20 \log \frac{\epsilon}{\epsilon_0} &= 20 \log \frac{a_0}{a} \\ &= \frac{20 \log (a_0/a_m)}{a/a_m} \\ &= N - N_0 \end{aligned}$$

where a_m is the amplification corresponding to zero inserted attenuation. This gives ϵ as $(N_0 - N)$ decibels above ϵ_0. It will, however, be more convenient to determine, on the above assumption with regard to noise voltage, the theoretical attenuation N_1 corresponding to a field intensity of 1 μV per metre. From the above,

$$N_1 = N_0 - 20 \log \epsilon_0$$

in which ϵ_0 is in microvolts per metre. The field strength ϵ corresponding to an attenuation N is then given directly as $(N - N_1)$ decibels above 1 μV per metre.

For example, from a plot of field intensity against N at 10 metres it was found that, for $N_0 = 50$, $\epsilon_0 = 1 \cdot 880$. Thus

$$\begin{aligned} N_1 &= 50 - 20 \log 1 \cdot 880 \\ &= 50 - 65 \cdot 50 \\ &= -15 \cdot 5 \text{ db.} \end{aligned}$$

The corresponding figures for 9, 8, and 7 metres were − 7·8, − 4·5, and − 4·6 db. respectively.

With such a set of constant numbers for the various wavelengths, the apparatus becomes virtually direct reading in field strengths.

For small values of N (less than about 10 db.) the standard output deflection will be partly due to "noise" —both internally generated and received from external sources. For example, at zero attenuation the output will be about 300 μA, even without additional background noise, due to the e.m.f. of thermal agitation of the first circuit, as modified by its connection to the aerial unit. Thus, in general, at small values of attenuation, the standard output K will fall, not to zero, but to some other value K_1 when the signal is removed. This residue must of course be allowed for in estimating the corresponding field-strength. The appropriate correction will obviously depend on the law of the second rectifier.

A linear rectifier, when excited by e.m.f.'s having r.m.s. values of E_1 and E_2, of different frequencies, gives an output proportional to $\sqrt{(E_1^2 + E_2^2)}$. Thus, for a linear rectifier, the correction required can be estimated by regarding the total standard output K as due to a signal e.m.f. E_s and a " noise " e.m.f. E_n That is,

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* See Bibliography, (5).

† *Ibid.*, (2).

$$K = k\sqrt{(E_s^2 + E_n^2)}$$

Also $$K_1 = k\sqrt{E_n^2} = kE_n$$

Therefore the response due to E_s alone is

$$kE_s = \sqrt{(K^2 - K_1^2)}$$

Thus if the calculated uncorrected field-strength is ϵ μV per metre, the corrected value ϵ_c is given by

$$\epsilon_c = \frac{\epsilon}{\sqrt{[1 - (K_1^2/K^2)]}}$$

or, in decibels,

$$N_c = N + 10 \log \left(1 - \frac{K_1^2}{K^2}\right)$$

Similarly, for a square-law rectifier,

$$\epsilon_c = \epsilon\sqrt{\left(1 - \frac{K_1}{K}\right)}$$

and $$N_c = N + 10 \log \left(1 - \frac{K_1}{K}\right)$$

In the present case, it was found by calibration that the second rectifier is linear down to about 50 μA output, and can therefore be regarded as linear over the working range.

(c) Analysis of Sensitivity Figures

It will be of some interest to analyse the recorded sensitivity figures. For example, at 7 metres, it is found that the theoretical field intensity for full standard output is 1·6 μV per metre. By calibration with a standard signal generator it is found that full standard output corresponds to 5·2 μV on the first grid.

Let k = frequency-conversion factor;
m_e = overall circuit magnification factor (i.e. ratio of secondary-circuit resonant p.d. to primary or aerial-circuit e.m.f.);
h_e = effective height of aerial coil
= $2\pi \times$ area$/\lambda$;
ϵ = field strength, in volts per metre;
and v = input voltage on first grid of intermediate-frequency amplifier.

Then $$\frac{v}{\epsilon} = kh_em_e$$

The aerial coil is approximately 33 cm. in diameter, giving a value for h_e of 7·7 cm. at a wavelength of 7 metres. From the above data,

$$\frac{v}{\epsilon} = 3{\cdot}25 \times 10^2$$

Thus $$km_e = 42{\cdot}2$$

It is shown in Appendix III that, for a directly connected intermediate-frequency circuit, k should not differ greatly from unity. In the present case, however, there is some degree of matching of the composite load consisting of the two tuned circuits and the transmission-line coupling to the much lower resistance of the diode, and from certain approximate measurements it appears that k is about 3, in which case m_e will be about 14.

It is shown in Appendix I that circuits having m values of m_1 and m_2, inductively coupled to the optimum degree as defined in Appendix I, should give an overall m (or magnification factor) equal to half the geometric mean of m_1 and m_2 times the square root of the ratio of the inductances L_1 and L_2. If it be assumed that an approximation to the optimum coupling has in fact been realized in the present case by the empirical adjustment used, then $(L_2/L_1)^{\frac{1}{2}}(m_1m_2)^{\frac{1}{2}}$ is about 28. This is a reasonable value, since the secondary circuit is a small coil of about $1\frac{1}{2}$ in. diameter, wound with solid wire, and is, moreover, damped by the diode.

It appears, therefore, that the numerical results obtained are in conformity with design estimates of the quantities involved at each stage.

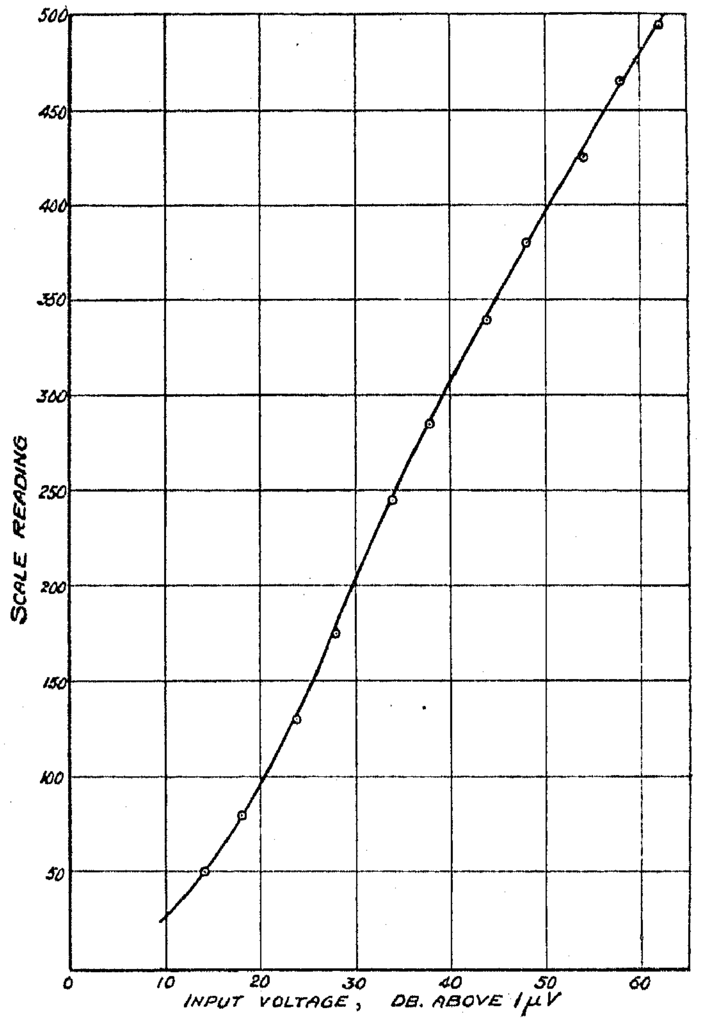

Fig. 7.—Output/input calibration with automatic gain-control.

(d) Operation with Automatic-Gain Control

It will be seen from Fig. 6 that the intermediate-frequency amplifier is provided with means for introducing automatic gain control if desired. This facility may be of some use in making continuous approximate surveys of field strength, e.g. in an aircraft or other moving vehicle. It is intended that the equipment should be used deflectionally in such cases, the automatic gain control enabling a wide range of field strength to be recorded within the limit of full-scale deflection.

The scale may be calibrated in terms of first-grid input-voltage by means of a standard-signal generator, coupled to the input circuit in such a way that the resulting input-circuit tuned impedance is reduced to the same value as is given by the coupling to the frequency-change unit (giving, therefore, the same input noise-voltage for initial adjustment). The actual voltage at the input grid can be determined in terms of the output voltage of the standard-signal generator by a separate measurement of the effect of the coupling, and the grid voltage may be related to field strength by means of the field-intensity calibration, as already described in (*b*) and (*c*), Section (5).

It was found that by switching to automatic gain control, and adjusting to a noise-voltage output of 50 (this initial deflection then being reduced to zero by means of the initial balancing control), a range of nearly 50 db. could be obtained from zero to full-scale deflection. The calibration curve is shown in Fig. 7. It is substantially linear in decibels, except at the lower end, where the co-existent noise voltage causes an apparent lowering of sensitivity.

The range can be extended by introducing a known attenuation by means of the intermediate-frequency attenuators. For example, at an input voltage of 62 db. above 1 μV the output was 495 μA. This was reduced to 130 μA by switching in 38 db. of attenuation in the intermediate-frequency circuit, which agrees exactly with the output for an input voltage of 62 — 38 (= 24) db. above 1 μV from the standard-signal generator.

(6) ACKNOWLEDGMENTS

The authors wish to acknowledge their indebtedness to Dr. J. S. McPetrie for the standard-field calibrations of the equipment described, and to Dr. R. L. Smith-Rose for criticism and advice. The work was carried out as part of the programme of the Radio Research Board, and is published by permission of the Department of Scientific and Industrial Research.

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APPENDIX I

For the inductively coupled system shown in Fig. 4, the transfer vector admittance i_2/e at frequency $\omega/(2\pi)$ is

$$\frac{i_2}{e} = -\frac{M\omega j}{Z_1 Z_2 + M^2\omega^2}$$

where $$Z = R + jX = R + j\left(\omega L - \frac{1}{\omega C}\right)$$

That is,

$$\frac{i_2}{e} = -\frac{M\omega j}{Z_1}\,\frac{1}{R_2 + \dfrac{M^2\omega^2 R_1}{|Z_1|^2} + j\left(X_2 - \dfrac{M^2\omega^2 X_1}{|Z_1|^2}\right)}$$

If M and X_1 are fixed, resonance with respect to X_2, i.e. with respect to the tuning capacitance C_2, is given by

$$X_2 = \frac{M^2\omega^2 X_1}{|Z_1|^2}$$

Similarly, if M and X_2 are fixed, resonance with respect to C_1 is given by

$$X_1 = \frac{M^2\omega^2 X_2}{|Z_2|^2}$$

For the first of these conditions,

$$\left|\frac{i_2}{e}\right|_{res.} = \frac{M\omega}{|Z_1|}\,\frac{1}{R_2 + \dfrac{M^2\omega^2 R_1}{|Z_1|^2}}$$

Regarding this as a function of X_1, it will be found by the usual process of differentiation that two optimum values for X_1 exist, i.e.

$$(a)\ X_1 = \pm R_1\sqrt{\left(\frac{M^2\omega^2}{R_1R_2}\right) - 1}\ \text{(maxima)}$$

$$(b)\ X_1 = 0$$

The first condition can only be fulfilled if the quantity under the radicle is positive. This is a condition of "excessive coupling," and corresponds to the well-known double-peaked resonance as a function of frequency. For this condition, the absolute optimum secondary current is given by

$$\left|\frac{i_2}{e}\right|_{res.} = \frac{1}{2}\,\frac{1}{\sqrt{(R_1R_2)}}$$

The second condition can be fulfilled for all values of M_1 and corresponds to a minimum with "excessive coupling" and a maximum with loose coupling. For this second condition

$$\left|\frac{i_2}{e}\right|_{res.} = \frac{1}{\sqrt{(R_1R_2)}}\,\frac{1}{\dfrac{\sqrt{(R_1R_2)}}{M\omega} + \dfrac{M\omega}{\sqrt{(R_1R_2)}}}$$

The factor involving $M\omega$ has a maximum value of $\frac{1}{2}$ when the two denominator terms are equal, i.e. when

$$M^2\omega^2 = R_1R_2$$

In this case the two optima (*a*) and (*b*) above merge into a single condition. In this condition $X_1 = X_2 = 0$. Putting

$$M^2 = k^2L_1L_2$$

$$k^2 = \frac{R_1}{\omega L_1}\cdot\frac{R_2}{\omega L_2}$$

or $$k = \sqrt{(p_1p_2)}$$

where p_1 and p_2 are the power factors of the two circuits.

Under conditions of optimum coupling, the output voltage v_2 is given by

$$\left|\frac{v_2}{e}\right| = \frac{M\omega + \omega L_2}{R_1R_2 + M^2\omega^2}$$
$$= \frac{1}{2}\sqrt{(m_1m_2)}\sqrt{\left(\frac{L_2}{L_1}\right)} = \frac{1}{2}\,\frac{L_2}{M}$$

where m is written for $\omega L/R$, i.e. the overall "m" is

$m/2$ in the case of optimum coupling of identical circuits, and half the geometric mean of the "m's" multiplied by the square root of the ratio of the inductances in the case of optimum coupling of circuits which differ.

APPENDIX II

Determination of Resistances and Coupling Factor for Loosely-Coupled Tuned Circuits

It is well known that the resistance and the "m" (i.e. $\omega L/R$) of a tuned circuit can be determined by measuring the capacitance-change δC corresponding to the $1/\sqrt{2}$ ordinates of the resonance curve, the formulae being

$$\frac{1}{2\omega}\frac{\delta C}{C^2} = R$$

and

$$m = \frac{\omega L}{R} = \frac{2C}{\delta C}$$

In the case of loosely coupled circuits, as illustrated in Fig. 4, assuming that the coupling is purely inductive and is such that

$$M\omega < \sqrt{(R_1 R_2)}$$

(i.e. a loose coupling, corresponding to a single-peaked resonance having an absolute maximum value when $X_1 = X_2 = 0$, as shown in Appendix I), the quantities $R_1 R_2$ and M can theoretically be determined by a somewhat similar process, as follows:—

(*a*) With C_1 constant at the resonance value corresponding to frequency $\omega/(2\pi)$, determine the total capacitance-change δC_2 corresponding to the $1/\sqrt{2}$ resonance ordinates of the secondary-circuit current (or voltage). Then, from the formula (see Appendix I):—

$$i_2 = -\frac{M\omega j}{(R_1 + jX_1)(R_2 + jX_2) + M^2\omega^2}$$

it can be shown that, with $X_1 = 0$,

$$\frac{1}{2\omega}\frac{\delta C_2}{C_2^2} = \frac{R_1 R_2 + M^2\omega^2}{R_1} = \alpha$$

(*b*) With C_2 constant at the resonance value, C_1 varied to give the $1/\sqrt{2}$ resonance values of secondary current (or voltage), we have

$$\frac{1}{2\omega}\cdot\frac{\delta C_1}{C_1^2} = \frac{R_1 R_2 + M^2\omega^2}{R_2} = \beta$$

(*c*) Remove or open-circuit the primary, and determine R_2 in the ordinary way, i.e.

$$\frac{1}{2\omega}\frac{\delta C_2'}{C_2^2} = R_2 = \gamma$$

Then R_2 is known, $R_1 = \beta\gamma/\alpha$, and $M^2\omega^2 = \beta\gamma[1 - (\gamma/\alpha)]$. Further,

$$\left|\frac{v_2}{e}\right|_{res.} = \frac{M\omega}{\omega C_2(R_1 R_2 + M^2\omega^2)}$$
$$= \frac{1}{\omega C_2}\sqrt{\left(\frac{\alpha - \gamma}{\alpha\beta\gamma}\right)}$$
$$= 2C_1\sqrt{\left[\frac{1}{\delta C_1}\left(\frac{1}{\delta C_2'} - \frac{1}{\delta C_2}\right)\right]}$$

Thus, considering the matter from the point of view of field-strength measurement, if v_2 is known, the effective e.m.f. e induced in the primary or aerial circuit is theoretically determinable by means of a set of three half-resonance measurements. The chief practical difficulties, at least at very high frequencies, would be (i) the determination of the tuning-condenser calibrations (which might, unless the condensers had been specially designed for ultra-short-wave operation, depend appreciably on frequency) and (ii) ensuring that the coupling is of the purely inductive character postulated. A further minor point is that the formulae are only exact in respect of resonances of the secondary current, whereas in practice the secondary resonance voltages would be measured. The difference in the resonance conditions is not, however, very material with reasonably good circuits.

The method was not actually adopted in the present instance, as the tuning condensers used were not suitable for the quantitative calibration involved.

APPENDIX III

Single-Phase and Biphase Diode Frequency-Conversion

Mathematical analysis.

The following is a simplified and approximate analysis of frequency conversion by means of diode rectification, by methods similar to those used by M. J. O. Strutt,* and is intended to give some qualitative insight into the process.

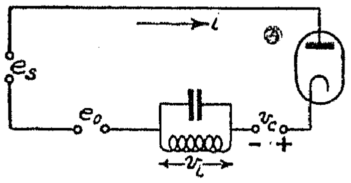

Fig. 8.—Diode frequency-change circuit [characteristic $i = f(v)$].

Consider the circuit shown in Fig. 8, where $e_s = \hat{e}_s \sin\omega_s t$ is a small signal voltage, $e_0 = \hat{e}_0 \sin\omega_0 t$ is a relatively large local-oscillator voltage, v_i is the output intermediate-frequency voltage, and v_c is a direct voltage somewhat smaller than e_0. (It may in practice be derived from a suitably by-passed resistance.) It will be assumed that the only significant external impedance in the diode circuit is the tuned intermediate-frequency (ω_i) circuit, having a dynamic resistance R_i.

Then

$$i = f(e_0 + e_s - v_c - v_i)$$

and, for sufficiently small values of e_s and v_i,

$$i = f(e_0 - v_c) + (e_s - v_i)f'(e_0 - v_c)$$

where

$$f'(v) = \partial f/\partial v$$

Bearing in mind the nature of $f(v)$, e_0, and v_c, it will be seen that $f'(e_0 - v_c)$ regarded as a function of time represents a series of pulses of the nature shown in Fig. 9, and as such can be expanded into a Fourier series. Thus

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* See Bibliography, (4).

$$i = f(e_0 \sin \omega_0 t - v_c)$$

$$+ (\hat{e}_s \sin \omega_s t - v_i)$$
$$\left[u_0 + u_1 \sin(\omega_0 t + \theta_1) + u_2 \sin(2\omega_0 t + \theta_2) + \ldots\right]$$

where u_0, u_1, u_2, etc., are constants.

In the above it is assumed that $f(v)$ is substantially independent of the frequency of v, which implies a suitable choice of diode at very high frequencies, and that the voltage e_s is maintained without effective internal impedance. Under these conditions the diode-capacitance current plays a negligible role as far as frequency conversion is concerned, and its omission in the above formulation is justified. Equating separately the intermediate-frequency components,

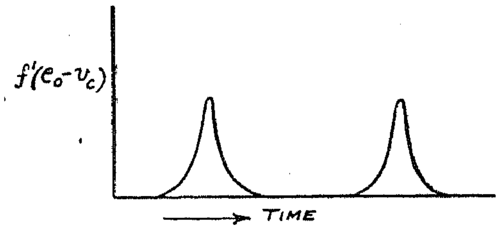

Fig. 9

$$i_i = -u_0 v_i + \hat{e}_s \cdot \frac{u_1}{2} \cos(\omega_i t + \theta_1)$$

where $$\omega_i = \omega_0 \sim \omega_s.$$

Putting $$v_i = R_i i_i,$$

$$v_i = \frac{R_i}{R_i + \dfrac{1}{u_0}} \cdot \frac{u_1}{2u_0} \cdot \hat{e}_s \cos(\omega_i t + \theta_1)$$

Thus, as far as the intermediate-frequency components are concerned, we have the equivalent circuit shown in Fig. 10.

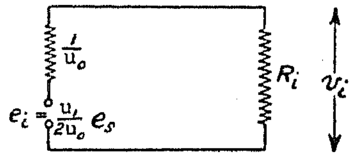

Fig. 10

The general character of the Fourier series must now be considered. Take, for example, the function shown in Fig. 11, where, for simplicity, the pulses are assumed to be the crests of sine waves of fundamental period T_0. It will be found that, for such a function,

$$u_0 = \frac{1}{\pi}(\sin \alpha - \alpha \cos \alpha)$$

$$u_1 = \frac{1}{\pi}\left(\alpha - \frac{1}{2} \sin 2\alpha\right)$$

$$u_2 = -\frac{1}{\pi}\left(\frac{1}{1\,.\,2} \sin \alpha - \frac{1}{2\,.\,3} \sin 3\alpha\right)$$

$$u_3 = -\frac{1}{\pi}\left(\frac{1}{2\,.\,3} \sin 2\alpha - \frac{1}{3\,.\,4} \sin 4\alpha\right)$$

$$u_4 = \frac{1}{\pi}\left(\frac{1}{3\,.\,4} \sin 3\alpha - \frac{1}{4\,.\,5} \sin 5\alpha\right)$$

where $\alpha = \pi\tau/T_0$.

Now if the pulses are narrow, i.e. if τ and α are small, so that $\sin \alpha$ can be replaced by $\alpha - (\alpha^3/6)$, then

$$u_0 = \frac{1}{\pi} \cdot \frac{\alpha^3}{3}$$

and $$u_1 = u_2 = u_3 = u_4, \text{ etc.} = \frac{2}{\pi}\frac{\alpha^3}{3} = 2u_0$$

This relationship is likely to hold whatever the shape of the pulse, provided it is sufficiently narrow. Moreover, even if it is an appreciable fraction of a period in width, u_1 will not differ greatly from $2u_0$. For example, if $\tau = T_0/2$,

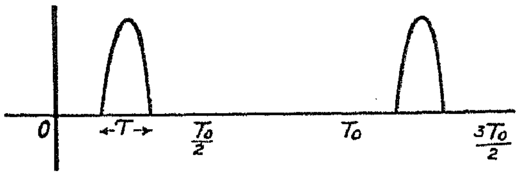

Fig. 11

Then $$\frac{u_1}{2} = u_0 = \frac{\pi}{4} \simeq 0{\cdot}8$$

Thus, in the above formula for v_i (the intermediate-frequency output voltage), $u_1/(2u_0)$ can be replaced by unity, and we have the simple relationship

$$\hat{v}_i = \frac{R_i}{R_i + \dfrac{1}{u_0}} \hat{e}_s$$

If, in addition, R_i is large compared with the effective internal intermediate-frequency resistance $1/u_0$,

$$\hat{v}_i = \hat{e}_s$$

This simplicity of theoretical relationship is of some value in relation to such quantitative applications as field-strength measurement. Even when the various postulated conditions cannot be very completely fulfilled, the use of a fairly large frequency-change oscillator voltage and an output circuit of high impedance will probably result in an overall conversion-ratio which does not differ much from unity.

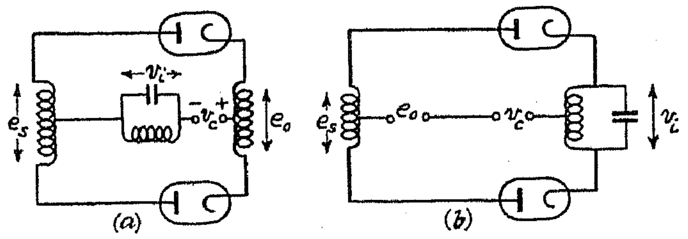

Fig. 12

If any form of direct coupling between the loop aerial and the rectifier circuit is used, the preservation of electrical symmetry will require the use of biphase or push-pull circuits, shown schematically in Fig. 12. The same simplified analysis for such cases gives for the intermediate-frequency output voltage a value just half that for the single-phase circuit.

Use of harmonics.

It has been shown that the fundamental and low-order harmonics of the diode-circuit current are of comparable magnitude. The local oscillator can therefore be adjusted to 2, 3, 4, etc., times the wavelength appropriate to frequency conversion by means of the fundamental, without appreciable loss of conversion ratio. This theoretical conclusion, which was pointed out by Strutt in the paper referred to, is substantially confirmed in practice. The conversion ratios corresponding to the second and third harmonics were found to be approximately equal, though a little less than that corresponding to the best conditions for the fundamental. The difference, which is not more than 3 or 4 db., is unimportant compared with the great advantages of harmonic operation, the chief of which is freedom from interaction between the local-oscillator and signal-frequency tuned circuits. At very short wavelengths there is the additional advantage that satisfactory oscillation is more readily obtainable at the longer wavelengths appropriate to the use of harmonics. A loop-aerial field-strength equipment for use at frequencies up to 300 Mc./sec. has already been constructed on these lines and has been found satisfactory.

Harmonic operation, biphase circuits.

The biphase circuits shown in Fig. 12 are appropriate for harmonic operation using odd harmonics. For even harmonics the oscillator voltage should be induced in the common lead in (*a*), and the output circuit should be in the common lead in (*b*).

THE DEVELOPMENT OF A SMALL VARIABLE AIR CONDENSER COMPENSATED FOR RAPID CHANGES OF TEMPERATURE *

By H. A. THOMAS, D.Sc., Member.

(*Paper first received* 10*th September, and in revised form* 16*th November*, 1938.)

SUMMARY

Various methods are discussed of obtaining compensation for temperature variation in air-dielectric condensers of small capacitance value suitable for radio-frequency purposes. It is found that appreciable air-gap spacing is necessary for high stability and a high degree of compensation. It is concluded that a parallel-plate system affords the most satisfactory means of achieving high stability.

A description is given of a variable air condenser of small dimensions and weight which has a capacitance range of 17–68 $\mu\mu$F and a high degree of temperature compensation; the overall dimensions are $2\frac{1}{2}$ in. × 2 in. and the weight is $5\frac{3}{4}$ oz. The temperature-coefficient of capacitance can be adjusted to have any value within the range zero to -120 parts in 1 million per degree Centigrade, and the compensation is effective at any rate of temperature variation not exceeding 5 deg. C. per minute.

When the system is used in association with a ceramic-former coil, its frequency coefficient can be made as low as 1 part in 1 million per deg. C., and the frequency stability over long periods with large temperature-change is of the order of 10 parts in 1 million.

(1) INTRODUCTION

In a recent paper† a description was given of a variable air condenser with adjustable compensation for temperature. The knowledge derived from a detailed study of the factors causing variation of capacitance‡ was used to make the compensation effective for comparatively rapid changes of temperature. The electrical performance was much superior to that obtained with previously-constructed temperature-compensated condensers. This condenser was, in fact, quite satisfactory for many purposes, but to meet other requirements it was decided to apply these same fundamental principles of compensation to a condenser of much smaller dimensions and reduced weight. With this object in view a study has been made of various means of constructing condensers of small bulk and high electrical stability.

(2) METHODS OF OBTAINING TEMPERATURE COMPENSATION IN SMALL AIR CONDENSERS

Methods alternative to a parallel-plate construction have been explored. Since the use of a very small air-gap enables an appreciable capacitance to be obtained with small dimensions, experiments have been made on two special condensers embodying unusually small air-gaps. The first condenser consisted of an eccentrically-mounted rotor of semi-cylindrical cross-section and an earthed cylindrical stator. The rotor was made of invar and was supported by spring-loaded balls, and the stator was made of brass. By turning the rotor through 180° the capacitance could be varied from 17 to 57 $\mu\mu$F. The performance of this condenser with temperature variation was cyclic, and the temperature-coefficients of capacitance at the maximum and minimum settings were -350 and -150 parts in 1 million per deg. C. respectively. A large negative value of the temperature-coefficient of capacitance was anticipated owing to the large fractional change of air-gap produced by the differential expansion of the rotor and stator.

This condenser is suitable as a trimmer with a large negative capacitance coefficient, but it is not satisfactory as a variable air condenser having an adjustable temperature-coefficient of capacitance for the following three reasons:—

(*a*) The temperature-coefficient of capacitance is large and negative, and its value cannot be varied over a wide range.

(*b*) Variation of the coefficient is accompanied necessarily by change of capacitance.

(*c*) Slight variations of the expansion coefficients of either the rotor or the stator cause considerable change of temperature-coefficient of capacitance, and accurate reproduction of the assembly is consequently difficult.

In the second experimental condenser an attempt was made to reduce the dependence of the coefficient of capacitance on the capacitance setting by making use of a method of compensation adopted previously.† This condenser consisted of a semi-cylindrical rotor mounted inside a hollow cylinder, steel balls being used as bearings; both rotor and stator were made of invar. Variation of capacitance was arranged by rotation of the earthed stator system with respect to the fixed rotor. The capacitance range was 20–66 $\mu\mu$F.

The temperature-compensating arrangement consisted of a brass tube fixed to the stator in such a manner that increase of temperature gave rise to expansion of this tube, and consequently to a relative axial movement between the stator and rotor. The dimensions were selected to give theoretically a negative coefficient of about 20 parts in 1 million per deg. C.

Tests on this condenser showed that the performance with temperature-change was cyclic but that the coefficient of capacitance had a large negative value. At the maximum and half capacitance settings the coefficients were − 125 and − 450 parts in 1 million per deg. C. respectively. This behaviour is due partly to the very small air-gap and partly to the difficulty of maintaining concentricity of the cylinders with angular rotation. Moreover, if the expansion coefficients of the rotor and stator are different by only 1 part in 1 million the temperature-coefficient of capacitance may be changed by 120 parts in 1 million per deg. C. Since the expansion coefficient of commercial invar may lie between − 2 and + 2 parts in 1 million per deg. C. it is apparent that the adoption of very small air-gaps may give rise to variations in capacitance coefficient between identically-constructed condensers of the order of 500 parts in 1 million per deg. C.

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* Official communication from the National Physical Laboratory.
Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 495.

† Thomas, H. A.: "A Variable Air Condenser with Adjustable Compensation for Temperature," *Journal I.E.E.*, 1937, vol. 81, p. 277, and *Proceedings of the Wireless Section*, 1937, vol. 12, p. 286.

‡ Thomas, H. A.: "The Electrical Stability of Condensers," *Journal I.E.E.*, 1936, vol. 79, p. 297, and *Proceedings of the Wireless Section*, 1936, vol. 11, p. 202.

† *Journal I.E.E.*, 1937, vol. 81, p. 277, and *Proceedings of the Wireless Section*, 1937, vol. 12, p. 286.

Tests on these two condensers have shown that the use of small air-gaps may give rise to considerable uncertainty of behaviour, owing to the fact that the expansion coefficients of the materials used are not known to a sufficient degree of accuracy, and normal methods of construction are insufficiently precise. If high electrical stability is desired, multiple air-gaps with appreciable spacing should be employed; there seems, in fact, to be no satisfactory alternative to a parallel-plate assembly.

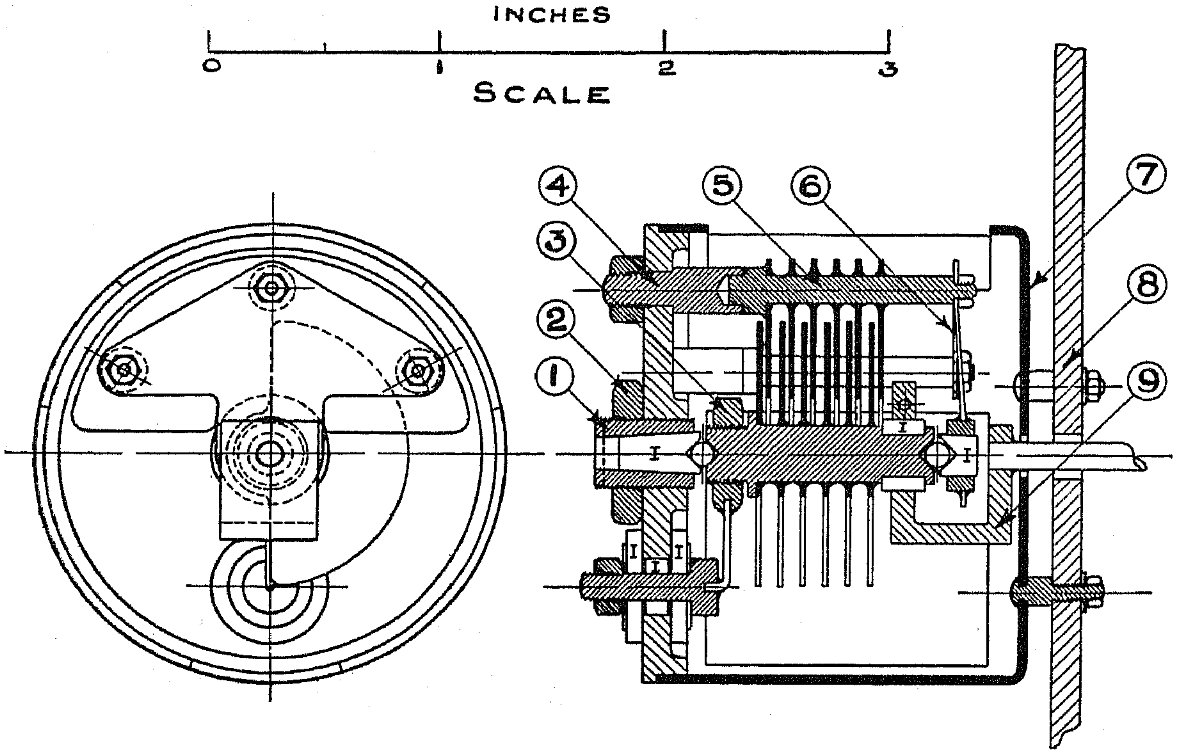

Fig. 1.—Assembly of temperature-compensated condenser.

(3) DESCRIPTION OF SMALL TEMPERATURE-COMPENSATED CONDENSER

The constructional details of this condenser are illustrated in Fig. 1. Since the temperature-compensating system functions by using the difference in expansion between different metals, it is necessary to construct both the stator and rotor assemblies of the same metal and to use another metal for the compensating assembly. Brass and iron are convenient metals for this purpose and, since a lesser degree of compensation is required if the stator and rotor plates are made of the material having the smallest expansion coefficient, these assemblies are made of iron. The stator and rotor plates are conventional in form and are soldered respectively to the three steel supporting rods (5) and the steel spindle. This spindle is located and insulated respectively by two steel balls and " Calit " plugs I, and is rotated by the insulated bracket (9) and drive shaft. Slight pressure is maintained on the locating balls by the spring plate (6).

Compensation for temperature-change is effected by three brass rods (4), which on expansion increase the air-gap on one side of the rotor plates and reduce the gap on the other side. The length of these rods is selected so that, when the air-gap on one side of the rotor plates is twice that on the other side, the differential expansion between the brass rods and steel shaft is such as to produce exact compensation for temperature. Adjustment of the temperature-coefficient of capacitance is provided by a screw (1); rotation of this screw produces an axial movement of the rotor assembly with respect to the stator; the adjusting screw may be locked in any position by means of the nut (2). The spring plate (6) serves the purpose of maintaining location of the rotor over an axial movement corresponding to the difference between the two extreme positions defined by contact between the rotor and stator assemblies. The negative temperature-coefficient of capacitance may be varied by adjustment of the rotor position relative to the stator plates.

In this particular condenser the rotor and stator vanes were made of iron and the compensating rods of brass, but it is not essential to adhere to this particular disposition. The stator and rotor assemblies can be made of brass or aluminium, in which case the compensating rods could be made of iron; if this modification is adopted a negative temperature-coefficient of capacitance is obtained when the rotor is moved as far as possible to the left (see Fig. 1), but the magnitude of the coefficient is less than that given with stator and rotor plates made of iron.

The stator is connected to the earthed frame and is enclosed in an aluminium case (7) provided with large ventilating holes; copper gauze is used to effect complete electrical screening. Connection to the rotor is provided by the sliding nut (3). The complete assembly is mounted behind the panel (8) by three pillars fixed to the aluminium case. All the metal parts are chromium-plated to give a smooth surface and prevent rusting. The overall length and diameter of the condenser are $2\frac{1}{2}$ in. and 2 in. respectively; the weight of the complete assembly is $5\frac{3}{4}$ oz. (160 g.). The capacitance range is 17–68 $\mu\mu$F for a 180° rotation of the driving spindle, and the power factor is about 5×10^{-4} at a frequency of 5 Mc./sec.

(4) DEGREE OF COMPENSATION OBTAINED

(a) Heating Tests on Condenser

The condenser was set up in the testing chamber used previously and tests were made to ascertain the possible range of adjustment of the temperature-coefficient of capacitance. It was found that any desired temperature-coefficient between zero and −120 parts in a million per deg. C. could be obtained by adjustment of the compensating screw. For a capacitance setting of 36 $\mu\mu$F the relationship between the temperature-coefficient of capacitance and the setting of this screw is shown in Fig. 2.

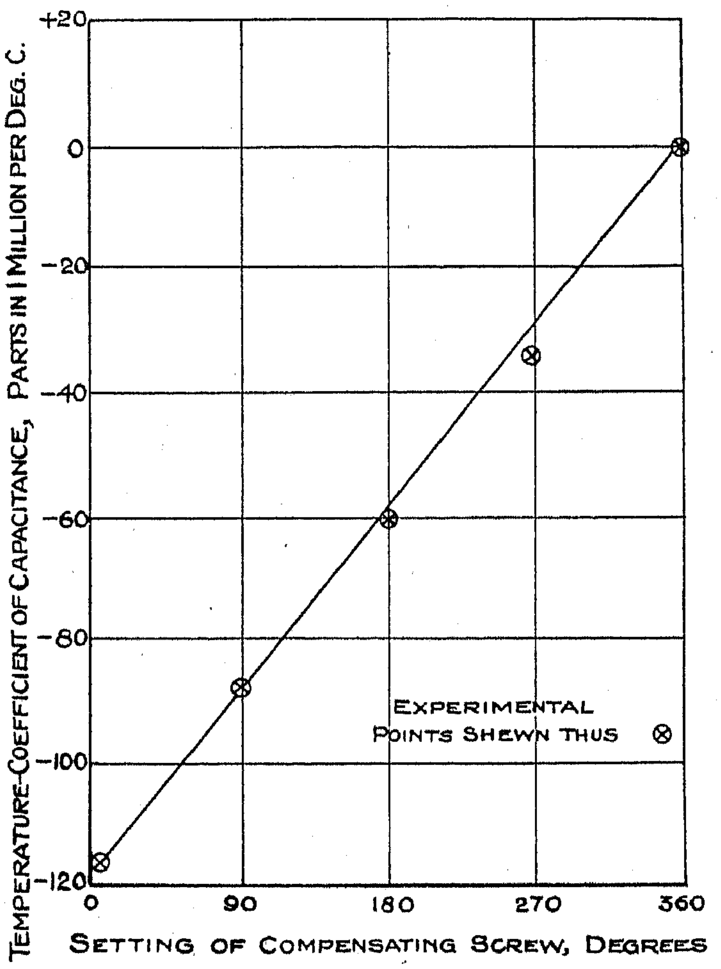

Fig. 2.—Relationship between capacitance coefficient and setting of compensating screw.

The compensating screw was adjusted to give a negative temperature-coefficient of about 18 parts in 1 million per deg. C. at a capacitance setting of 68 $\mu\mu$F, and tests were undertaken to ascertain the electrical performance under varying conditions of temperature-change. The nature of the capacitance variation can best be illustrated by reference to Fig. 3, which shows the effect of heating the condenser at a rate of about 5 deg. C. per minute. It is seen that at this rate of heating the capacitance variation followed very nearly the temperature-change, the temperature-coefficient of capacitance being −18 parts in 1 million per deg. C. At slower rates of heating no transient effects of any kind were observed.

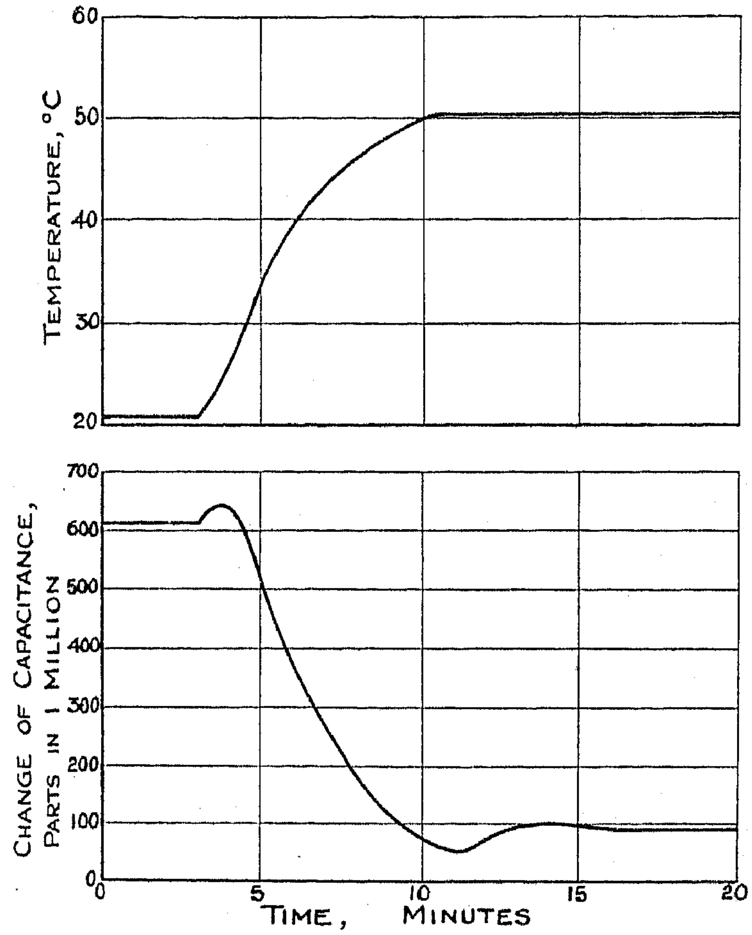

Fig. 3.—Example of rapid heating.

At the same setting of the compensating screw the temperature-coefficient of capacitance was measured at various capacitance settings; the results are shown in Table 1.

Table 1

Variation of Temperature-coefficient of Capacitance with Capacitance Setting

Capacitance, $\mu\mu$F	Coefficient of capacitance, parts in 1 million per deg. C.
25	− 97
36	− 59
55	− 30
68	− 18

In this Table the temperature coefficients have been deduced from the observed frequency variations on the assumption that the valve and circuit capacitance of 17 $\mu\mu$F is constant.

The results demonstrate that considerable variation of temperature coefficient occurs with setting of the rotor; this is probably due to the fact that the rotor and stator vanes are not exactly parallel. This variation of coefficient with capacitance setting is obviously not important when frequency stabilization is required over only a narrow frequency band. For applications in which a high degree of stabilization is required over a wide frequency band it is probable that this variation of capacitance coefficient with rotor setting could be reduced appreciably by improvement of the assembly with respect to the parallelism of the rotor and stator plates. It must, however, be pointed out that, even if the temperature coefficient were made independent of the setting, the other capacitances of the circuit (i.e. valve, coil, and lead capacitances) would necessarily limit the range of frequency over which a high degree of stabilization could be expected.

(b) Heating Tests on a Temperature-compensated Oscillator

The condenser was now associated with a ceramic-former coil of inductance $11\cdot4\,\mu$H to form a closed circuit in which oscillation was maintained by a dynatron valve system; the mechanical and electrical properties of the coil have been described in a previous paper.* The capacitance was adjusted so that the oscillation frequency was $6\cdot5$ Mc./sec. At this frequency the temperature-coefficient of inductance of the coil was $+28$ parts in 1 million per deg. C.

The complete oscillator was put into an oven and the compensating screw was adjusted until the frequency coefficient of the oscillation circuit was very small. The temperature was increased regularly from 20° to 50° C. in 12 hours and was then decreased to 20° C. during the next 12 hours. This thermal cycle was repeated for 7 consecutive days, during which observation was made frequently of the oscillation frequency. It was noticed that the frequency-changes followed exactly the temperature variations and that the frequency coefficient of the oscillator was about $+1\cdot5$ parts in 1 million per deg. C. The total drift in frequency over this period of 1 week was less than 10 parts in 1 million.

(5) CONCLUSION

Observation of the performance of this condenser shows that a very high degree of frequency stabilization can be attained with a ceramic-former coil in association with a condenser of special construction. The condenser necessary for this purpose can be made quite small in both dimensions and weight if the required capacitance range is not very large. Such a condenser can be designed to respond to variations of temperature as rapid as 5 deg. C. per minute, this rate of temperature-change being greater than that obtained in nearly all practical applications.

(6) ACKNOWLEDGMENTS

The work described in this paper was carried out as part of the programme of the Radio Research Board and is published by permission of the Department of Scientific and Industrial Research. The author is indebted to Mr. R. G. Chalmers for assistance in the experimental work, and to Mr. F. G. Murfitt for the construction of the condenser.

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* H. A. Thomas: "The Dependence on Frequency of the Temperature-Coefficient of Inductance of Coils," *Journal I.E.E.*, 1939, vol. 84, p. 101, and *Proceedings of the Wireless Section*, 1939, vol. 14, p. 19.

OUTPUT AND EFFICIENCY OF THE SPLIT-ANODE MAGNETRON OSCILLATING IN THE DYNATRON REGIME[*]

By A. F. HARVEY, B.Sc.(Eng.), Graduate.[†]

(*Paper first received* 21*st September, and in revised form* 15*th November*, 1938.)

SUMMARY

An account is given of the performance of the split-anode magnetron when working at frequencies such that transit time can be neglected. An accurate and convenient method is described of determining this performance from the measurement of valve resistance at power frequencies. A series of results for one valve are included; these show the general influence of the various operating conditions.

The anode modulation and frequency stability of the valve are discussed, while some measurements at radio frequency are given and the agreement of the two methods is found to be good.

(1) INTRODUCTION

The split-anode magnetron consists of a central filament with two hemi-cylinders external to it. In operation an axial magnetic field is applied and it is well known that as the field strength is increased the electron paths for a given anode voltage become more curved until finally, at a value called the "critical field," they just graze the anode and return to the filament; the anode current then becomes small. If the filament diameter is small compared with the anode diameter d, then Hull showed that this value of field is given by

$$H = \sqrt{\left[\frac{181\,V}{d^2}\right]}$$

where H is measured in gauss, V in volts, and d in cm., and is independent of space-charge conditions.

Under this condition of "cut-off" the valve resistance of each half anode can become negative over a certain range, and if a tuned circuit of sufficient dynamic impedance is connected between the anodes, as shown in Fig. 12, sustained oscillations are possible. From similarity with the negative-resistance triode and tetrode these are commonly referred to as "dynatron oscillations." The mechanism by which this effect is produced in the valve is partly understood[‡] and will not be discussed further.

Provided the frequency is less than, say, 10 Mc./sec.[§] the effects of electron transit time can be neglected and the valve performance can be predicted by measurement made at power frequencies, or even from the static characteristics.

The valve used in the experiments (Osram type CW.10) had an anode diameter of 1 cm. and a length of 2 cm., with a rated dissipation of 50 W. An iron-cored electromagnet was used for the field, the air-gap being 3·8 cm. long and the diameter of the pole-pieces 10 cm. The field strength was determined from the magnetizing current, a previous calibration having been made, by a fluxmeter, care being taken to reduce the error due to magnetic hysteresis.

(2) THE STATIC CHARACTERISTICS

An account of these has been given by E. C. S. Megaw[‡] and as an insight to the dynatron oscillation they will be briefly dealt with. Fig. 1 shows the currents taken by each anode for three different values of magnetic field (including zero). It is seen that with the magnetic field the lower-voltage electrode takes the larger current except when the amplitude of alternating voltage is large. Fig. 2 gives the cycle of currents under a given set of operating conditions. The full-line curve represents the power component of current in the tuned circuit, and Fourier analysis of this enables the output of the valve to be determined. Since the anode current is sensibly cut off when the valve is not oscillating, it is clear that the battery supplies power by virtue of the rectified current which results from the oscillation, and this is shown on the same diagram. The valve performance, however, can be predicted more readily from other experimental curves which will now be described.

(3) METHOD OF DETERMINATION OF VALVE RESISTANCE

It is known that for a tuned circuit of low power factor the fluctuations of potential of the anode to which it is connected are nearly simple harmonic. The valve current, however, is complex and the fundamental component only supplies the power absorbed in the external circuit and load. This gives rise to a quantity, the fundamental component of valve resistance, which in the case of the magnetron, as we have shown above, is negative.

It will be understood that the amplitude of oscillation which can be maintained in a circuit must have that value for which the fundamental component of resistance of the valve is equal and opposite to the dynamic resistance of the circuit. Hence a low-frequency measure-

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[*] Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 683.

[†] Engineering Laboratory, Oxford University.

[‡] See, for example, E. W. B. GILL and K. G. BRITTON: *Journal I.E.E.*, 1936, vol. 78, p. 461 and *Proceedings of the Wireless Section*, 1936, vol. 11, p. 127; also K. POSTHUMOUS: *Wireless Engineer*, 1935, vol. 12, p. 126.

[§] In the absence of magnetic field, the transit time is given approximately by the equation

$$T = \frac{2\cdot 5}{10^2}\cdot\frac{d}{\sqrt{V}}\text{ microsec.}$$

The critical field increases T by a factor of about 4; hence if $V = 100$, $d = 0\cdot 5$, T is of the order 1/200 microsec.

[‡] *Journal I.E.E.*, 1933, vol. 72, p. 326, and *Proceedings of the Wireless Section*, 1933, vol. 8, p. 72.

ment of this fundamental component will enable the performance of the magnetron at any frequency appreciably less than the inverse of the transit time of electrons to be predicted.

The fundamental component of resistance can be measured at power frequencies by a dynamometer, for if the complex anode current passes through the suspended coil and the field due to the fixed coils alternates simple-harmonically at the fundamental frequency of the anode current, the resulting couple is due only to that fundamental component. A fuller description of this method as applied to triodes has been given previously.*

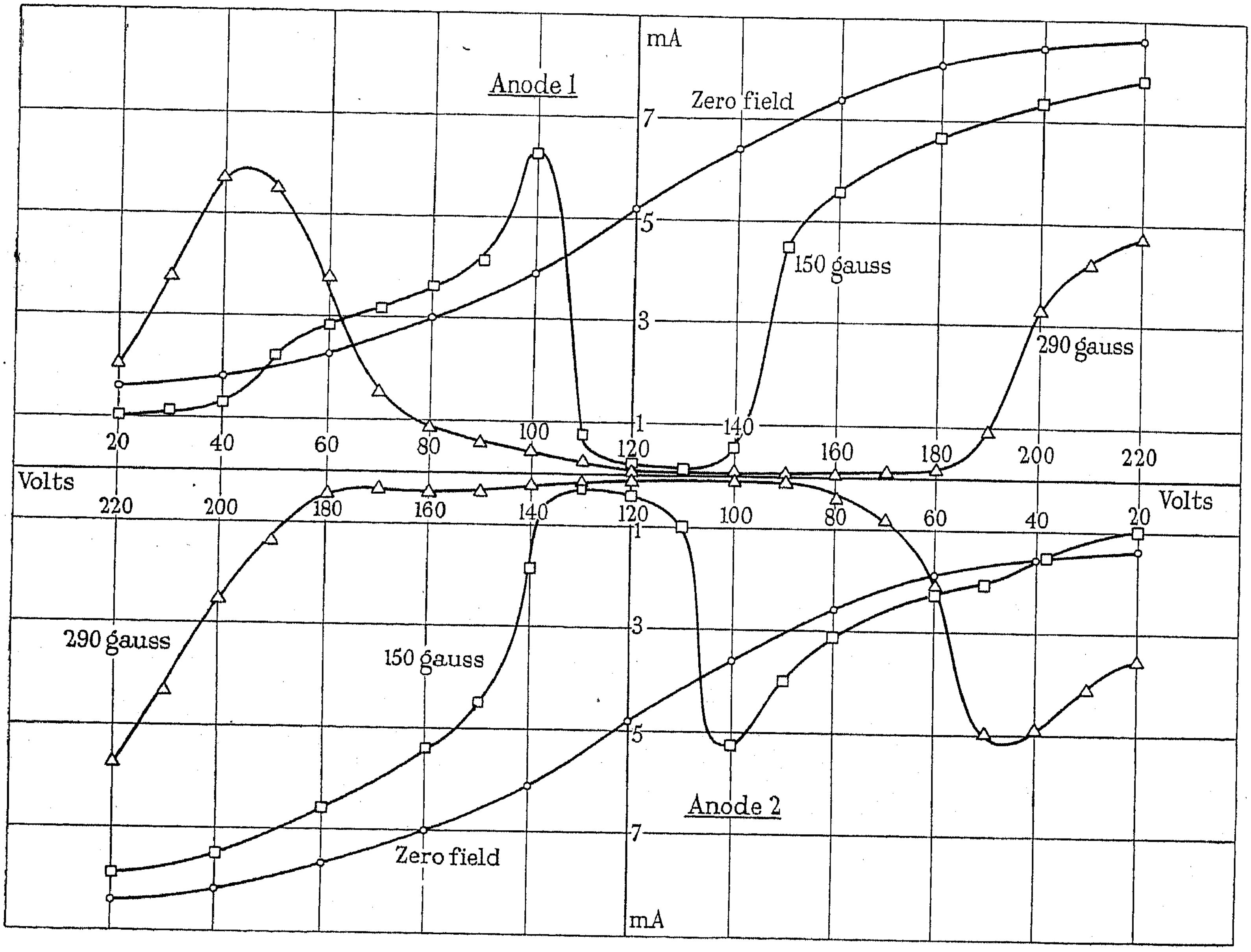

Fig. 1.—Static curves of magnetron. Filament emission 10 mA.

The circuit diagram of the measuring system is shown in Fig. 3: it is self-explanatory and need not be described in detail. The H.T. supply was connected to the centre tap in the suspended coil, and the suspension strips were used to connect the two ends of the coil, through identical windings of a transformer, to the anodes. The moving coil had an inductance of 18 mH and a resistance of 60 ohms, and the fixed coils had an inductance of 12 μH: thus the respective reactances were small compared with the resistances in series with them.

The negative resistance of the valve at any assigned amplitude was determined by matching the observed deflection with that which resulted when the valve was replaced by a standard non-inductive resistance. (The deflections were given the same sense by reversing the leads to the moving coil.) The sensitivity was such that resistances up to 0·25 MΩ could be measured with applied potentials of the order of 100 V (r.m.s.).

The accuracy of the method was verified by making a Fourier analysis of oscillograms of the anode current. An example taken under conditions differing only slightly from those stated in Fig. 2 is given in Fig. 4, the resistance value measured as above being almost identical with the value revealed by analysis. The error in employing a sinusoidal anode voltage is small for coils of reasonable power factor, as shown in the Appendix.

(4) EXPERIMENTAL RESULTS

The relation between the fundamental component of negative resistance and the r.m.s. value of the sinusoidal fluctuations of anode potential are shown in Fig. 5(*a*). The resistance is high for small amplitudes and reaches a minimum when the amplitude increases, and finally again becomes large and ultimately positive. If the rejector resistance of the tuned circuit is less than this minimum value then oscillation will not be maintained. Since the resistance is high for small amplitudes, it is only possible to start the oscillation by temporarily

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* D. A. Bell: *Journal I.E.E.*, 1935, vol. 76, p. 415, and *Proceedings of the Wireless Section*, 1935, vol. 10, p. 97; E. B. Moullin: *Wireless Engineer*, 1938, vol. 15, p. 371.

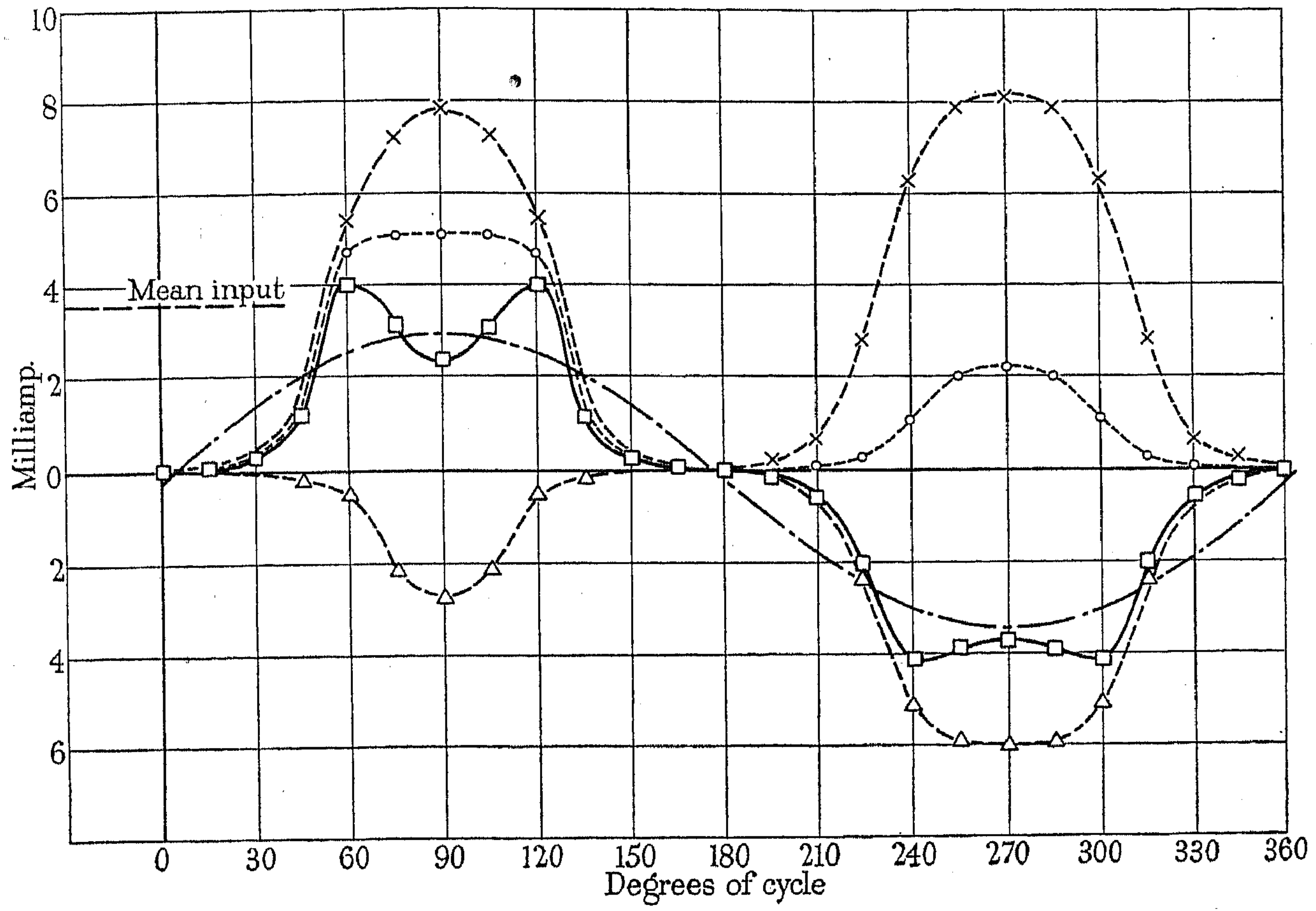

Fig. 2.—Cycle of currents deduced from Fig. 1:—

Anode 1 o - - - - - o
Anode 2 Δ – – – – Δ
Output ($A_1 - A_2$) □——□
D.C. input ($A_1 + A_2$) × – – – – ×
Output (fundamental) ——•——
Total a.c. voltage 110 (r.m.s.).
Field strength, 290 gauss.

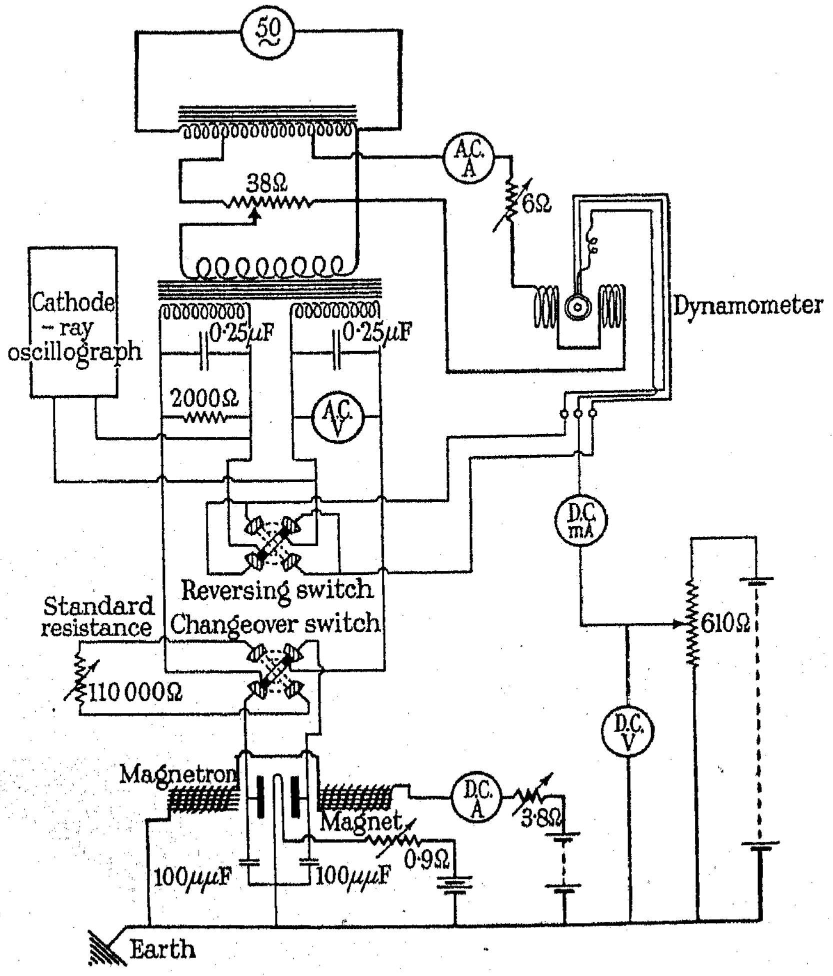

Fig. 3.—Diagram of connections used for measurement of magnetron impedance.

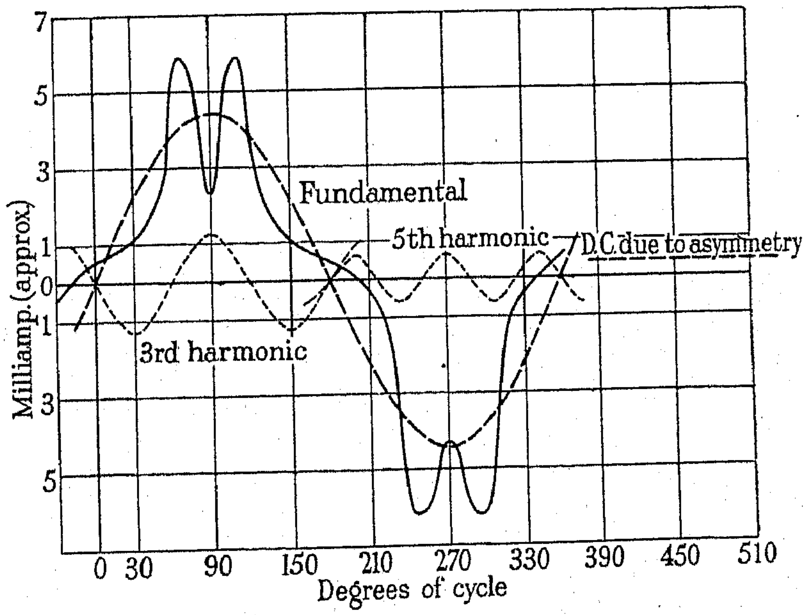

Fig. 4.—Wave-form of magnetron, deduced from oscillogram.

increasing the circuit resistance or reducing the magnetic field. For any circuit resistance higher than the minimum it is seen that two amplitudes are possible, the smaller one being unstable.

Ignoring the very small power dissipated in the circuit by the higher-harmonic components of current, the output would be V^2/R, where R is the negative resistance appropriate to the amplitude V. Fig. 5(*b*) shows this output; the curves rise to a maximum which occurs very near to the value of V giving minimum resistance.

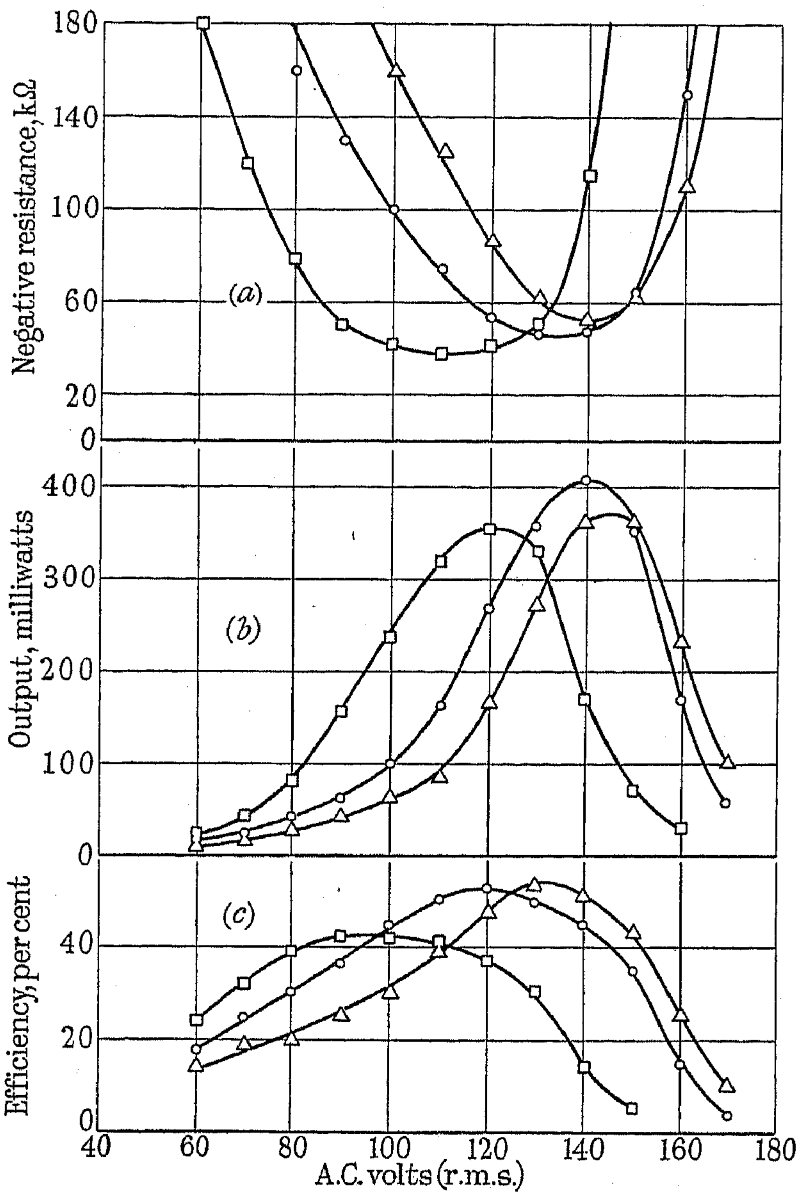

Fig. 5

Anode voltage	120
Filament emission	50mA
Field 220 gauss	□——□
Field 290 gauss	○——○
Field 345 gauss	△——△

In practice this optimum resistance value would be obtained by two tapping points on the coil. Measurement of the rectified anode current enables the efficiency to be calculated, and this is shown in Fig. 5(*c*). Comparison of these sets of curves shows that the amplitude for maximum efficiency is nearly that for maximum output and that there is a best value for the magnetic field which in general is about twice the critical field. These optimum conditions are examined below.

Fig. 6(*a*) shows the manner in which the load resistance for maximum output is related to the voltage of the H.T. supply, for different filament emissions. The field strength for maximum output increases in a nearly parabolic manner with anode voltage, as shown in Fig. 6(*b*), the effect of filament emission being small.

Figs. 7(*a*) and 7(*b*) show the relation between maximum output and efficiency, respectively, with supply voltage, and it will be seen that it is important to have as large a filament emission as possible. It was not possible to use supply voltages up to the maximum for this particular valve (1 000 V), but Fig. 8 gives a relation between output and supply voltage and filament emission which can be used as a guide. It is seen that this relation is best given by the equation

$$W = 0{\cdot}55 \times 10^{-4} \times V^{3/2} \times I^{1/2}$$

where W is in watts, V in volts, and I in mA.

(5a) EFFECT OF TILTING THE MAGNETIC FIELD

The previous experiments were made with the direction of the magnetic field coincident with the axis of the electrodes. Fig. 9(*a*) shows the effect on the output of inclining the electrode axis to the field. The upper and lower curves relate to a constant field strength, and show the relation between maximum output and angle of tilt: it may be seen that in these conditions the output may be increased or decreased by tilting the electrodes. These changes, however, are not due to adjustment of the axial component of the field strength as such (thus cos 6° = 0·9945), but are related to pronounced changes in the foot of the cut-off curve. The middle curve of Fig. 9(*a*) relates to a test in which the field strength was adjusted to give maximum output at various angles of tilt. Fig. 9(*b*) shows the corresponding curve of efficiency in the three conditions of test used in Fig. 9(*a*). The efficiency is small at large angles of tilt, because then the current cannot be cut off by the field in the static state. Examination of the wave-form of current showed that the harmonic content in certain cases was reduced slightly by tilting the field: this is probably associated with the small rise of efficiency and output.

(5b) MODULATION

The behaviour of the valve under conditions of anode modulation for use as a radio-telephone transmitter is given in Fig. 10 for a constant value of magnetic field and different circuit impedances. On the rising portion of the curve only a certain depth of modulation can be used without causing the oscillations to cease, but the linearity is good. The falling portion enables nearly full modulation to be obtained but the linearity is not good, the efficiency is poor due to the employment of high anode voltage at a comparatively low field strength, and, as will be seen below, considerable frequency modulation is present.

(5c) FREQUENCY STABILITY

In the absence of changes due to temperature, etc., in the values of the circuit parameters the frequency stability depends mainly on the harmonic content of the anode current, and the analysis given for the dynatron* can be applied to the magnetron directly.

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* E. B. Moullin: *Journal I.E.E.*, 1933, vol. 73, p. 186, and *Proceedings of the Wireless Section*, 1933, vol. 8, p. 177.

Some measurements (made at radio frequency as described later) are given in Fig. 11 (*a*, *b*, and *c*), the latter being of importance in showing the variation in carrier frequency under anode modulation. Provided the valve is operated near its best conditions the frequency changes as shown by the full lines are small, but become large when the conditions are such that the output is small and the current wave-form very distorted. Comparison with Fig. 10 shows that the value of anode voltage for maximum output corresponds to the position of minimum frequency.

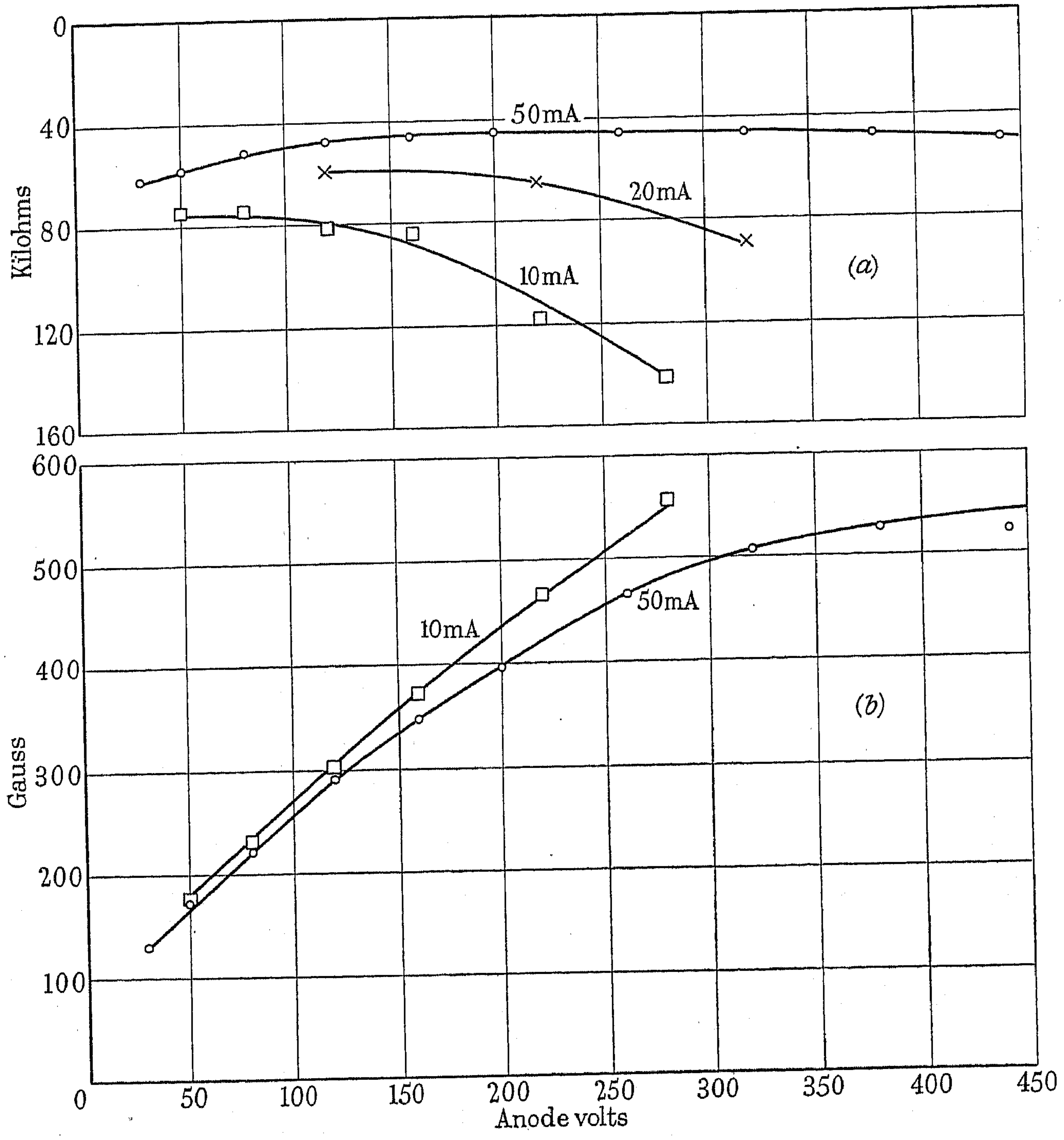

Fig. 6

(*a*) Variation of optimum load impedance with anode voltage.

Total filament emission 10 mA □——□
Total filament emission 50 mA ○——○
Total filament emission 20 mA ×——×

(*b*) Variation of optimum field strength with anode voltage.

Total filament emission 10 mA □——□
Total filament emission 50 mA ○——○

(6) RADIO-FREQUENCY MEASUREMENTS

To check the validity of the method of predicting the performance in terms of the fundamental component of negative resistance appropriate to a sinusoidal fluctuation of anode voltage, the output was measured when the system was acting as a self-contained generator.

A frequency of about 200 kc./sec. was chosen because it was convenient for measurement, far removed from the conditions of tests at 50 cycles per sec. and also from the region of resonance oscillations. The circuit diagram is drawn in Fig. 12, which also shows the values of the various circuit components. It was necessary to connect the two anodes to the cathode by condensers of 100 $\mu\mu$F capacitance as shown, to prevent short-wave oscillations. The high-frequency resistance of the coil and condenser (when connected to the cold magnetron) was measured in the frequency range 170–290 kc./sec.: in this range the power factor was constant to within

± 2 % and had a mean value of 0·6 % (i.e. $Q = 166$). The current through the condenser was measured by a thermal ammeter, or deduced from the reading of an electrostatic voltmeter when the resistance of the thermal ammeter (1·02 ohms) was objectionable. The load impedance was varied by means of a 10-ohm calibrated resistance box valid at 200 kc./sec.

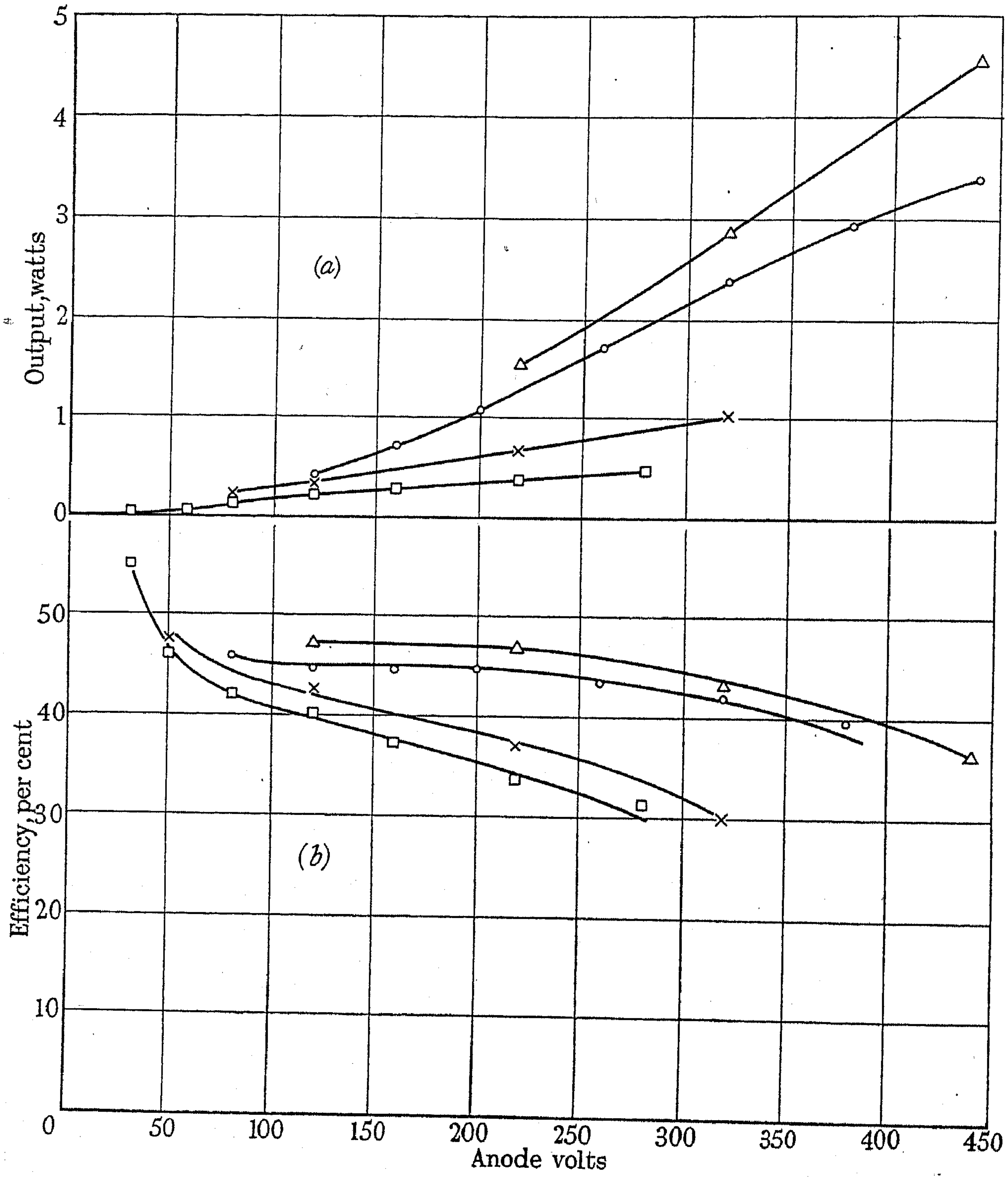

Fig. 7

(a) Variation of maximum output with anode voltage.
(b) Variation of efficiency at maximum output with anode voltage.

Total emission 80 mA △——△
Total emission 50 mA ○——○
Total emission 20 mA ×——×
Total emission 10 mA □——□

Measurements of output were made in a variety of operating conditions, shown collected in Tables 1 and 2. The last two columns of Table 1 show respectively the output in mW measured when oscillating at 200 kc./sec. and predicted from the tests at 50 cycles per sec. such as are recorded in Figs. 5, 6, etc., and the agreement is satisfactory. In Table 2 the ammeter and series resistance was removed and then the circuit had a rejector resistance of 63 kΩ. The output voltage was measured for various supply voltages, and the measured and predicted values are shown for comparison in the 6th and 7th columns of Table 2: here the agreement is usually closer than 1 % and well within the accuracy of measurement. The figures in the last two columns of Table 2 show for comparison the measured and predicted input;

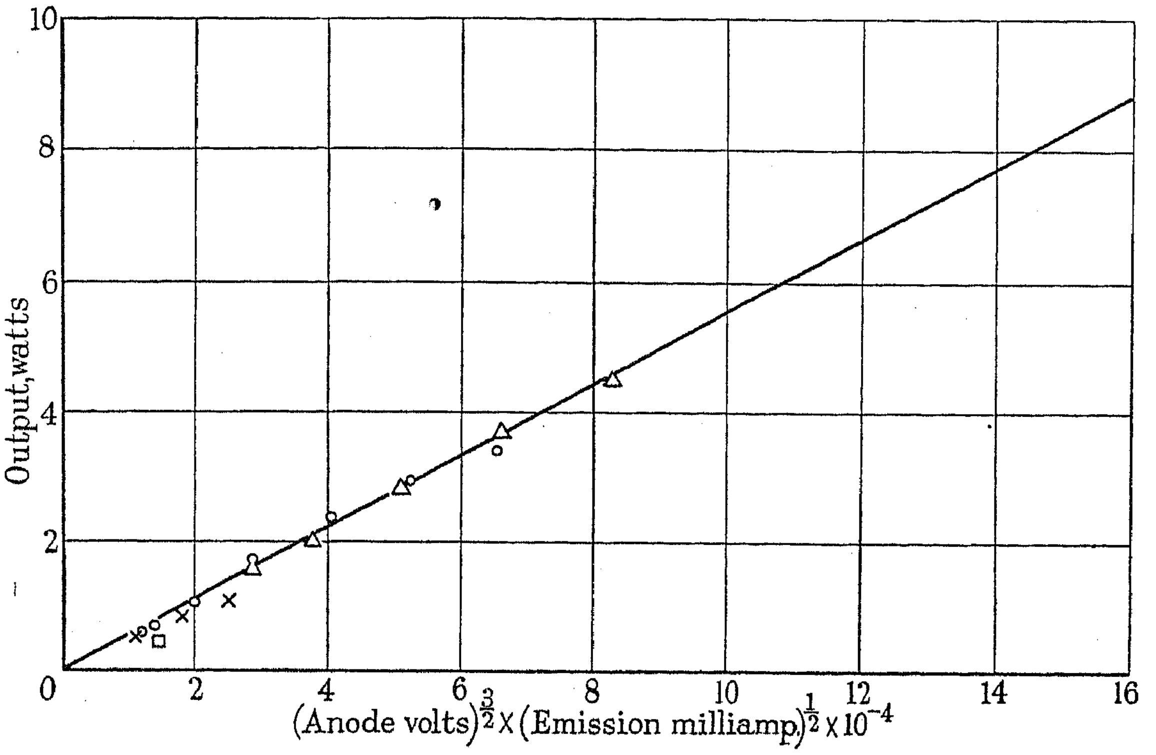

Fig. 8.—Variation of output with anode voltage and filament emission of magnetron.

Emission 80 mA △ Emission 50 mA ○ Emission 20 mA × Emission 10 mA □

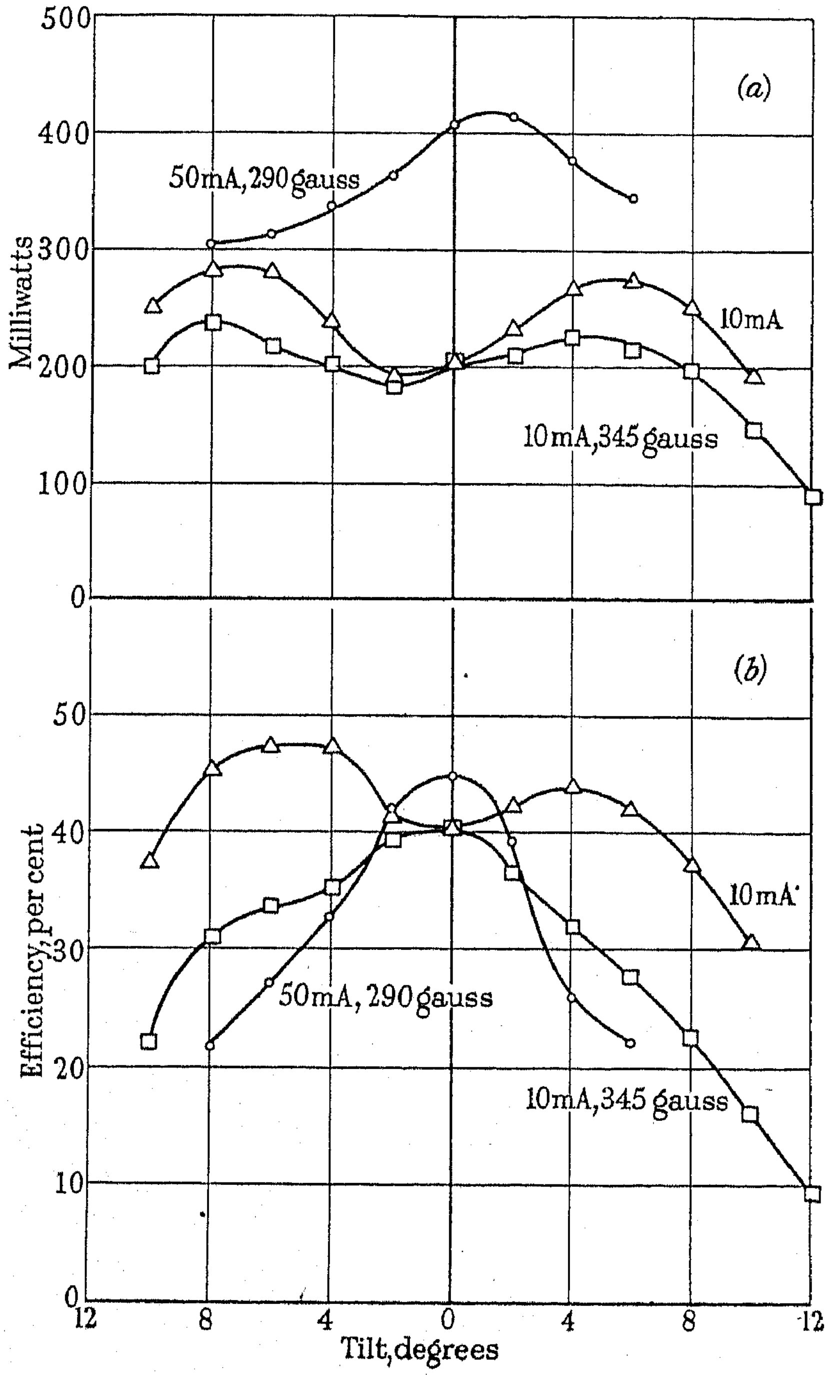

Fig. 9

(*a*) Variation of power output with tilt of field, at constant anode voltage (120) and saturation emission.
(*b*) Variation of efficiency with tilt of field, at constant anode voltage (120) and saturation emission.

With constant field and optimum load impedence ○——○
With constant field and optimum load impedence □——□
With optimum field and optimum load impedance △——△

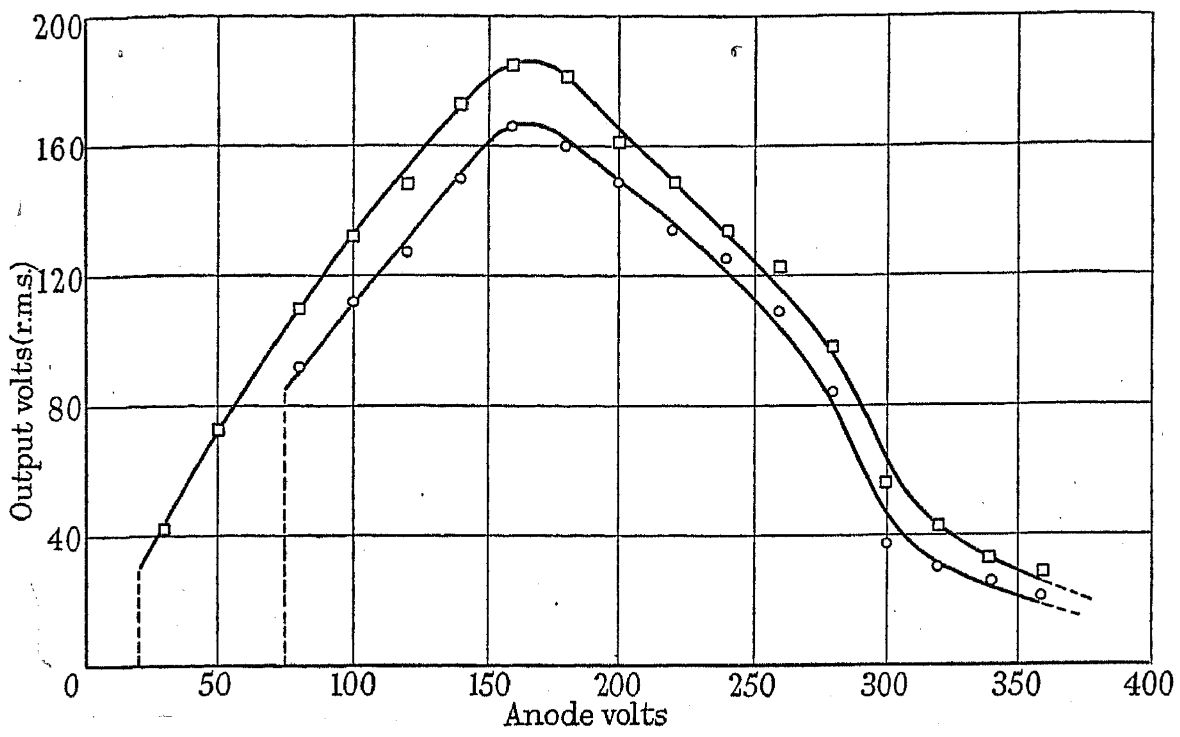

Fig. 10.—Variation of output voltage with anode voltage.

Field strength 230 gauss.
Filament emission 50 mA.
Load impedance 50 kΩ. ○——○
Load impedance 200 kΩ. □——□

Table 1

Anode volts	Field	Saturation emission	Angle of tilt	Series resistance	R.F. current	Output	Value at 50 cycles	Discrepancy
	gauss	mA	°	ohms	mA	mW	mW	%
220	230	50	0	6·19	286	508	526	+ 3·5
160	230	50	0	5·20	333	580	605	+ 4·3
140	230	50	0	4·20	307	398	438	+ 10
120	230	50	0	3·22	294	342	337	— 1·4
120	290	50	+ 5°	4·20	250	264	290	+ 10
120	255	50	0	3·22	303	296	282	— 4·7
120	255	80	0	3·22	307	304	310	+ 2·0

Table 2

Anode volts	Field	Saturation emission	Angle of tilt	Dynamic impedance	R.F. volts	Volts at 50 cycles	D.C. input	Input at 50 cycles
	gauss	mA		ohms			mA	mA
220	345	50	0	63 000	246	244	—	—
200	345	50	0	63 000	233	232	—	—
180	345	50	0	63 000	216	214	—	—
160	345	50	0	63 000	198	204	—	—
140	345	50	0	63 000	174	176	—	—
120	345	50	0	63 000	150	152	6·50	7·02
100	345	50	0	63 000	125	123	—	—
120	290	50	0	63 000	150	149	8·20	8·35
120	220	50	0	63 000	135	134	9·40	9·50

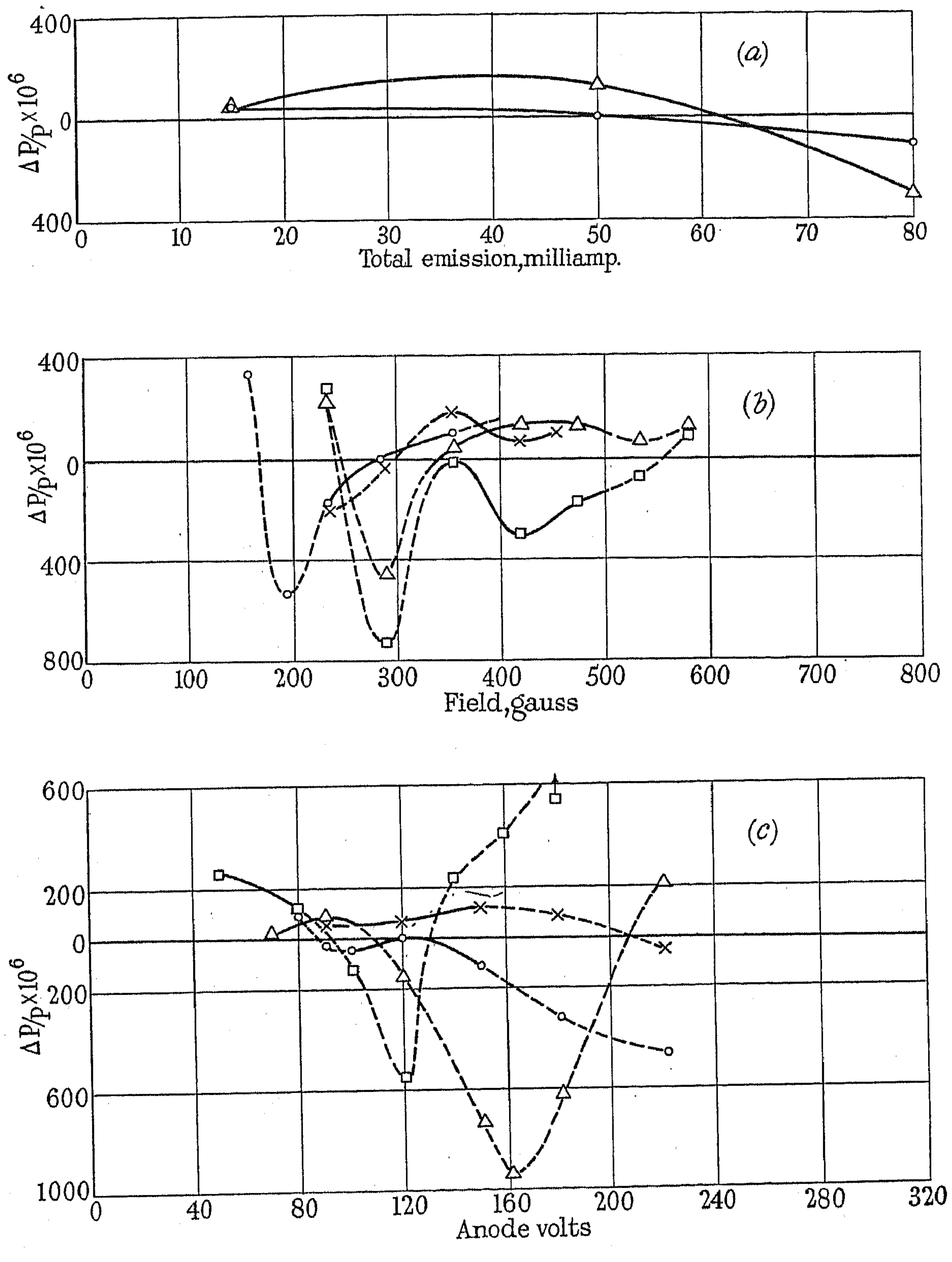

Fig. 11.—Variation of frequency with operating conditions. Mean frequency 200 kc./sec.; load impedance 63 kΩ.

(*a*) 220 volts, 420 gauss △———△
120 volts, 290 gauss ○———○

(*b*) 120 volts, 50 mA ○———○
220 volts, 50 mA △———△
220 volts, 15 mA ×———×
220 volts, 80 mA □———□

(*c*) 235 gauss, 50 mA △———△
290 gauss, 15 mA ×———×
290 gauss, 50 mA ○———○
195 gauss, 50 mA □———□

here again the agreement is satisfactory. Thus these measurements perhaps suffice to establish the validity of the method of prediction.

(7) CONCLUSION

The paper has demonstrated that the performance of the split-anode magnetron can be predicted with good accuracy by measurements at power frequencies made as described. The method is much more convenient than deduction from static curves, since to obtain information as to the best performance under various conditions a large number of curves would have been necessary and a Fourier analysis made in each case.

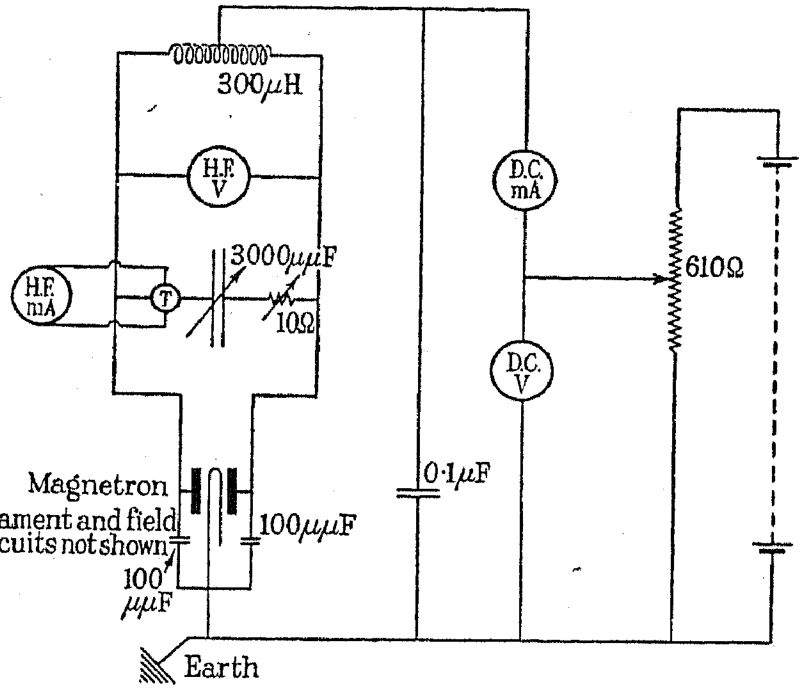

Fig. 12.—Diagram of connections for radio-frequency measurements.

The experimental curves given in the case of the particular valve enable the best operating conditions and circuit design to be chosen for any required output. In addition, they are capable of predicting accurately the performance at frequencies up to that at which electron transit time commences to play a part.

ACKNOWLEDGMENTS

The author's thanks are due to Mr. E. B. Moullin, M.A., Member, who has supervised the work, which was carried out in the Engineering Laboratory of the University of Oxford.

APPENDIX

The experiments recorded in this paper relate to simple-harmonic fluctuations of anode potential. This condition cannot obtain in a self-maintained generator since the resulting anode current is complex and must flow through the oscillatory circuit. However, the properties of this circuit are such that the harmonic components of voltage are much less than the corresponding components of current. Thus if V_1, V_3, V_5, etc., are the voltage components due to currents I_1, I_3, I_5, etc., flowing through a rejector circuit tuned to I_1 and having a power factor F, then it is well known that

$$\frac{V_3}{V_1} = \frac{3}{8}\cdot\frac{I_3}{I_1}F, \quad \frac{V_5}{V_1} = \frac{5}{24}\cdot\frac{I_5}{I_1}F, \quad \text{etc.}$$

Fourier analysis of the typical anode current represented by Fig. 4 shows that $I_3/I_1 = 0{\cdot}28$ and $I_5/I_1 = 0{\cdot}14$: hence it follows that if F is less than $0{\cdot}1$ no harmonic component of voltage is as great as 1 %. Hence it may be seen that the assumption of simple-harmonic voltage fluctuation is a very close approximation to the conditions obtaining in practice in a self-maintained generator. To examine whether the experimental results are sensitive to small changes in the wave-form of anode voltage the test corresponding to that recorded in Fig. 4 was performed with a peaked wave-form of voltage containing 3rd- and 5th-harmonic components of approximately 5 % and 3 % respectively. It was found that such change of wave-form resulted in the output being decreased by 11 %, from 409 to 363 mW. Thus it is clear that the method described for predicting the performance of a magnetron is sensibly correct even when the oscillatory circuit has a power factor much greater than would usually be tolerated in practice.

QUASI-STABLE FREQUENCY-DIVIDING CIRCUITS*

By R. L. FORTESCUE, M.A., Graduate.†

(*Paper first received* 1*st July, and in revised form* 25*th November*, 1938.)

SUMMARY

The paper describes some of the physical mechanism involved in frequency-division by valve circuits; requirements which must be satisfied for such division to occur without free self-oscillation being derived. A practical circuit for this purpose is developed from this theory, and the form in which it has actually been used described; together with the results of some experiments on stability carried out on dividers of this type. In conclusion, some more complex systems which might form subjects for future practical investigation are indicated schematically.

It is well known that an oscillatory system supplied with energy at frequency nf can be synchronized to oscillate at an exact sub-multiple frequency f, when f lies between certain frequency limits. Any oscillatory circuit synchronized in this way provides a frequency divider; and most practical demultiplying circuits, including the multivibrator, are of this type.

There are, however, disadvantages in the use of self-oscillatory circuits for frequency division, for under certain conditions they may fail to synchronize, and so produce an output which is not at the correct sub-multiple frequency. When this difficulty is serious, special circuits which can only oscillate in the synchronized state are of value. Because they do not show the self-oscillatory property associated with unstable systems, circuits of this type can be described as "quasi-stable," while the fact that they always synchronize requires the added qualification that they are "interlocked."

The first quasi-stable interlocked circuits are apparently those due to Sterky‡ and Longo.§ They are mainly of use for division by 2, and consist of back-coupled balanced modulator circuits. A number of more general circuits appear to exist, however, and it is proposed (*a*) to describe the physical requirements for quasi-stable division; (*b*) to trace the design of a practical divider using an interlocked quasi-stable circuit which has been used for division by 3, 4, 5, and 6; and (*c*) to indicate some other circuits of the same type which might provide useful and interesting subjects for research.

(*a*) Quasi-stable frequency-dividing circuits are only extreme cases of synchronized oscillators, and the principle of their physical operation can be derived from consideration of the simplified valve oscillator shown in Fig. 1.|| A 1 : 1 transformer with no leakage is assumed, and also a high-impedance valve (e.g. pentode), whose anode current, for practical load resistances, is independent of the anode voltage-swing. These simplifications, while not invalidating the physical argument, avoid obscuring the principles by subsidiary effects.

It is proposed to examine the effect of harmonics in the grid-voltage wave-form on the energy fed back by the valve, and its reaction winding L_2, to the oscillatory circuit L_1,C; and to show that the amplitude and phase of the fundamental term fed back in this way depend upon the amplitude and phase of these harmonics. The reaction can therefore be regarded as composed of two parts, (1) that due directly to the fundamental in the grid voltage, and (2) that produced as a result of harmonics in the grid voltage. A quasi-stable dividing system results from the absence of the first of these parts.

Consider a grid-voltage wave-form

$$v_g = v\left[\sin \omega t + k_n \sin (n\omega t + \phi_n)\right]$$

(taking only one harmonic, for simplicity)

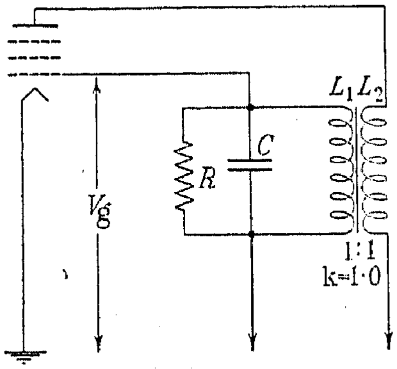

Fig. 1.—Simplified valve oscillator circuit.

The valve anode current, assumed independent of anode voltage-change, may be written

$$i_a = av_g + bv_g^2 + cv_g^3 + \ldots$$

Substituting the expression assumed for v_g, a Fourier series for i_a can be derived. The fundamental term of this series will consist of $av \sin \omega t$, plus a component from such products as

$$\left(v^{n-1} \sin^{n-1} \omega t\right) \left[vk_n \sin (n\omega t + \phi_n)\right]$$

which only appear in virtue of the presence of the harmonic in the grid circuit, and the non-linear terms in the valve characteristic.‡ An example, with second harmonic (i.e. $n = 2$), and valve with square-law characteristic, is given in Appendix 2 to show more exactly how this extra term arises. It is easily seen that harmonics

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* Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 693.

† Formerly with the British Broadcasting Corporation.

‡ *Ericsson Technics*, vol. 2, p. 31; and *Proceedings of the Institute of Radio Engineers*, 1937, vol. 25, p. 1153.

§ *L'Onde Électrique*, 1934, vol. 13, p. 372.

|| See Appendix 1 for additional notes on physical theory involved.

‡ There will also be terms from the odd powers of v_g which do not involve harmonics; these limit the amplitude to which oscillations build up.

can "intermodulate" in this way between themselves as well as with the fundamental, and the general effect may be expressed by writing the fundamental term of the anode-current expression

$$I_0 = av \sin \omega t + f_1(v, a, b, \ldots k_n, \ldots \phi_n, \ldots \text{etc.}) \sin \omega t + f_2(v, a, b, \ldots k_n, \ldots \phi_n, \ldots \text{etc.}) \cos \omega t$$

where $av \sin \omega t$ is the "linear" term which is used in simplified linear oscillator theory, and the rest is an added "intermodulation" term including a reactive component, arising from the non-linear valve characteristic which occurs in practice (and also from any harmonics which may be introduced into the circuit from external sources).

It is now clear that the magnitude and phase of the fundamental term of the reaction-coil current are, in part, dependent upon the harmonics in the grid circuit. In Fig. 1, when there is a steady maintained oscillation the grid voltage must be equal to the product of the L,C,R circuit impedance and the anode current. We have therefore an identity, at fundamental frequency,

$$v \sin \omega t \equiv Z\,[av \sin \omega t + f_1 \sin \omega t + f_2 \cos \omega t]$$

where f_1 and f_2 are the amplitudes of the intermodulation terms depending on the harmonics. For this identity to hold, the frequency must be such that the circuit L,C,R has a phase angle given by

$$\text{arc tan}\,\frac{-f_2}{av + f_1}$$

This is, in general, not zero, and the free frequency is therefore, in general, not that at which the circuit is resistive, $1/(2\pi\sqrt{LC})$; as would have been deduced by linear theory.

A practical oscillator, therefore, self-oscillates at a frequency shifted from its theoretical value as calculated by linear analysis. As this frequency-shift depends upon the harmonic intermodulation, introduction of an extra harmonic from an external source can vary it. Thus an oscillator which has a fundamental of 1 000 c./sec. when 3 000 c./sec. is injected from an external source in series with its grid, may have a frequency of, say, 1 200 c./sec. if this harmonic is removed; while its frequency calculated by linear analysis is

$$\frac{1}{2\pi\sqrt{LC}} = 1\,300 \text{ c./sec.}$$

This steady oscillation at a sub-multiple of an injected harmonic is the action which enables frequency-division to be obtained. For, in the case just considered, an oscillator of 1 200 c./sec. free oscillation *can** oscillate steadily at exactly a third of 3 000 c./sec. when this harmonic is injected into the circuit to "lock" it. The effect is due directly to the valve non-linearity and injected harmonic; and the greater the ratio of "intermodulation feedback" to "direct feedback," the more this injected harmonic can control the frequency of oscillation.

It now appears that the synchronization effect is not fundamentally dependent upon the presence of "direct feedback" at all, and, therefore, that if the "intermodulation feedback" is increased, when normal reaction is reduced, the total feedback can still be sufficient to maintain oscillation. If this process is taken far enough, the direct reaction becomes insufficient for self-oscillation in the absence of the injected harmonic, and a quasi-stable system is obtained.

The physical requirement for a quasi-stable circuit is, therefore, (*a*) a low direct feedback that is not sufficient to cause free self-oscillation, and (*b*) a sufficient feedback due to intermodulation with the injected harmonic, to maintain oscillation when such a harmonic is injected from an external source, in spite of this reduced direct reaction. A reduction of direct reaction without loss of "intermodulation" effect may be obtained in practical form either by using the balanced circuits of Serky and Longo or more generally by using frequency-discriminating circuits. A practical frequency-divider using the latter is shown in Fig. 2; its operation can now be considered, and it can be shown to be interlocked as well as quasi-stable.

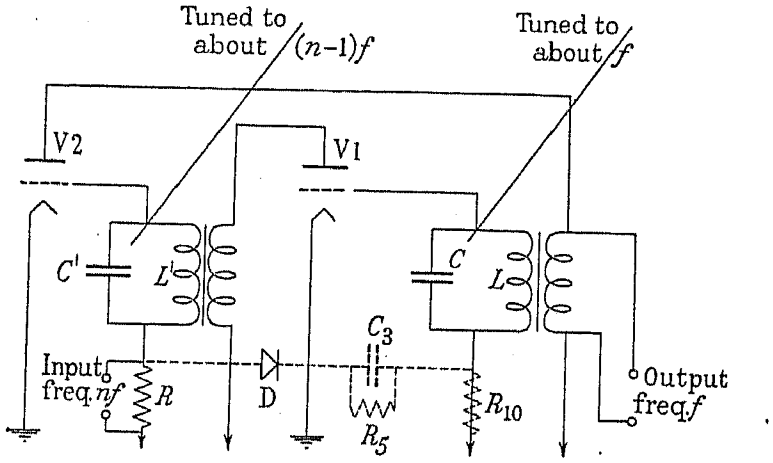

Fig. 2.—Simplified quasi-stable frequency-dividing circuit.

(*b*) Commencing from the valve oscillator of Fig. 1, it is required to obtain a circuit with reduced relative amount of direct reaction compared with that due to intermodulation with injected energy. To achieve this, the relative amplitude of the intermodulating harmonics, compared with the fundamental, in the grid-voltage wave-form of the reacting valve must be increased. It is therefore necessary to separate this reaction valve from the main oscillatory L,C circuit, tuned to the fundamental, across which the wave-form is necessarily good. By using a valve under conditions favourable to non-linearity as a separating stage, a large relative amplitude of harmonics is obtained in its anode current, despite this essentially good wave-form of the grid voltage across the tuned circuit L,C. Further, by tuning the anode circuit of this separator stage to a selected harmonic, the voltage on the grid of the reacting valve at that harmonic frequency can be made considerably greater than the voltage at fundamental frequency. The circuit of Fig. 2 is then arrived at, and it represents the simplest form of quasi-stable divider.

The operation of the circuit of Fig. 2 under working conditions can now be summarized:—

(1) The main oscillatory circuit L,C is tuned to about the required output frequency f. It is maintained in

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* See Appendix 3.

oscillation by means of the reaction coil in the anode circuit of V_2.

(2) The auxiliary circuit L',C' is tuned to a harmonic, $(n \pm 1)f$. When the main circuit L,C is oscillating at f, the separator valve V_1 acts as a harmonic generator and drives this auxiliary circuit L',C' at frequency $(n \pm 1)f$. V_1 also magnifies the fundamental of frequency f, but as the anode circuit is out of tune as far as this frequency is concerned, the circuit impedance is low and the magnification is very small; so that the voltage handed on to the grid of V_2 is almost entirely of the harmonic frequency $(n \pm 1)f$, at which the circuit is tuned.

(3) The resistance in the grid circuit of V_2, across the source of energy whose frequency is to be divided, results in this grid also being driven at frequency nf.

(4) The valve V_2, acting as a modulator with grid voltages at frequency nf and $(n \pm 1)f$, gives an anode current (which passes through the reaction coil) including the frequencies $nf \pm (n - 1)f$; which are the "sidebands" of nf when modulated with $(n \pm 1)f$. One of these terms is of frequency f, and the current component of this frequency in the reaction coil maintains the oscillation in the main circuit L,C when the loop gain is sufficient.

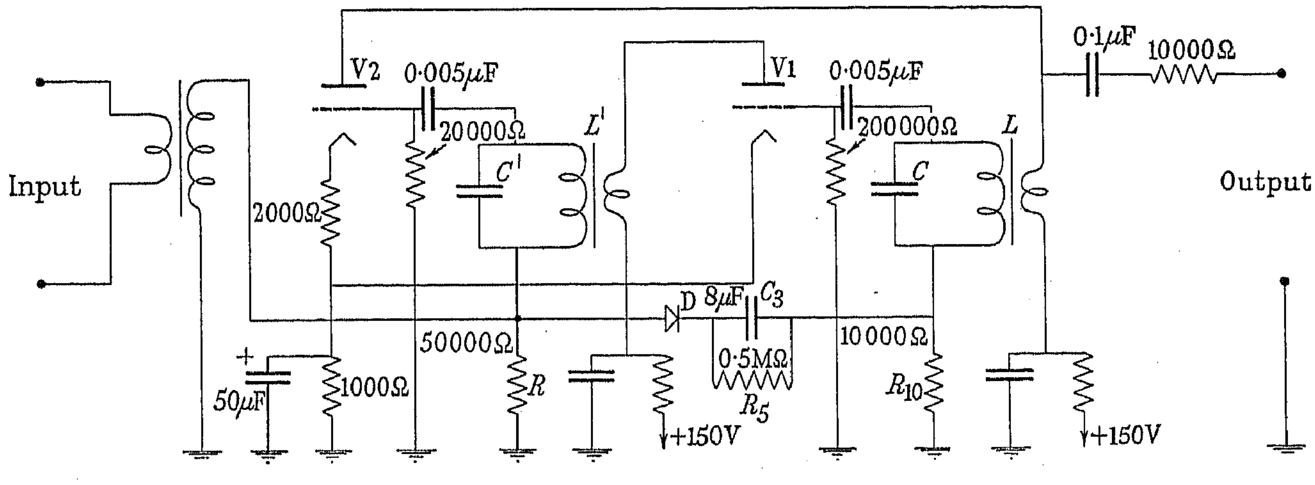


Fig. 3.—Practical quasi-stable circuit.

V and V_2 are Marconi H30 valves. L and L′ are dust-cored transformers with anode winding 8 mH and grid winding 130 mH. For division from 3 150 to 1 050 c./sec. the capacitances were $C = 0{\cdot}12\ \mu F$ and $C' = 0{\cdot}03\ \mu F$.

This action can only occur when the grid of V_2 is driven at frequency nf from the input; so the circuit is quasi-stable, direct reaction round the loop being avoided by the tuned circuits being sharply tuned to widely differing frequencies. The oscillation is also interlocked, for suppose it started at some frequency, different from the exact sub-multiple, $f(1 + \delta)$. The harmonic generator gives $(n \pm 1)f(1 + \delta)$, and the lower sideband when modulated with nf is :—

$$nf - (n \pm 1)(1 + \delta)f = \pm f(1 + \delta) - n\delta f$$

This is not of the same frequency as the original $f(1 + \delta)$, and therefore cannot maintain the oscillation.

Owing to the finite attenuation of the circuit L',C' at frequency f, and the circuit L,C at frequencies nf, $(n \pm 1)f$, direct reaction, though greatly reduced, is not completely eliminated, and it is found in practice that the remnant thereof should be arranged to be in correct sense (at least in simple circuits), to avoid increasing the apparent damping of the tuned circuits by negative reaction effect. Some care is also required in practice in adjusting the overall gain so that, while this remanent reaction does not cause self-oscillation, the intermodulation term is great enough to do so. The most satisfactory practical circuit actually tried is of the form shown in Fig. 3, where it is arranged for division by 3. The operation is identical with that just described with $n = 3$ and V_1 operating as a frequency doubler.

A fundamental difficulty arises in starting the oscillations in these circuits, on account of the output from the frequency multiplier V_1 (Fig. 2) of $(n \pm 1)$th harmonic being proportional to the $(n \pm 1)$th power of its input voltage, when this is small. The loop gain round the feedback path is therefore proportional to $v^{[(n\pm1)-1]}$, where v is the voltage across the main oscillatory circuit L,C. At the low levels of circuit noise, etc., which normally build up an oscillation, this gain may be less than unity, and therefore such building-up will not occur. The first experimental circuits of this kind were therefore started by whistling the required note into a microphone across the oscillatory circuit, which was interesting but laborious. A simple automatic form of starting circuit is shown in Figs. 2 and 3. The principle relies on the fact that by driving V_1 (Fig. 2), initially with frequency nf as well as f, it will act as a modulator and give, as sidebands, an output of $(n \pm 1)f$ as required [see (2) in col. 1], which is proportional to v instead of to $v^{(n\pm1)}$. The loop gain is then independent of level, and circuit noise, etc., is sufficient to build up an oscillation. But because the system is not interlocked, although it is quasi-stable, such an oscillation is not necessarily synchronized. Once the amplitude of the starting oscillation has built up, however, V_1 gives enough gain as a harmonic generator to maintain the interlocked oscillation, and the extra drive on V_1 can be reduced until this interlocked oscillation is the only form possible. This "extra drive" is supplied across a resistance R_{10} in the grid circuit of V_1, as shown dotted in Fig. 2, and the condenser and leak arrangement C_3,R_3, charged by a rectifier D, gives a gradual reduction of the "starting voltage" on V_1 as the condenser charges up. The divider is thus made to interlock after a time determined by the product of C_3 and R_5, as well as being self-starting on application of input (provided that the input has been removed for long enough for the condenser C_3 to discharge, a point which can be avoided, at the expense of some stability, by leaving some " residual " starting drive on V_1 at all times).

Owing to the use of tuned circuits, the operation of the divider of Fig. 3 is confined to a narrow frequency band near their resonant frequencies. The efficiency of the modulator stage V_2 depends also upon the input level of frequency $3f$ applied to it, no oscillation occurring if this is too high or too low. A range of levels and frequencies over which oscillation will occur is thus obtained, and a plot of this range is given in Fig. 4, for a particular divider with rather sharp tuning. The effect of variations of battery voltage is also shown, to give an indication of the improved stability of such a system compared with multivibrators or high L/C oscillators.

There are a variety of circuits of the quasi-stable interlocked type which appear to offer interesting possibilities for investigation. Fig. 5(*a*) shows a logical development of the circuit of Fig. 3, in which band-pass filters are used instead of simple tuned circuits. A preliminary and rather short experiment on this circuit indicated that the increase in frequency band was not as much as expected, however. Fig. 5(*b*) is the same circuit with high- and low-pass filters, while the arrangement of Fig. 5(*c*) shows a band elimination circuit. Starting provides something of a problem in these types on account of there being no sharply tuned circuit across which an oscillation can easily build up. It is also difficult in practice to obtain sufficient amplitude from a harmonic generator to maintain oscillation unless it is connected to a tuned circuit favouring the required harmonic.

Finally, Fig. 5(*d*) shows the first arrangement in its natural form for dividing by very high factors. An oscillator at nearly the correct frequency, 1 kc./sec., drives a succession of multipliers to give, say, 27 kc./sec. This output is fed, together with the input of 28 kc./sec., to a modulator. The output of this modulator is 1 kc./sec., and the auxiliary oscillator can now be faded slowly out, the oscillation being maintained by this 1 kc./sec. output from the modulator. Division by 28 should then be obtained without instability.

ACKNOWLEDGMENT

Acknowledgment is due to the British Broadcasting Corporation for permission to publish the paper, which is a result of work done in their Research Department.

APPENDIX 1

The physical description in terms of " reaction " here given is, of course, only one way of regarding the synchronization effect which occurs in all non-linear oscillatory systems. Amongst others, a point of view which has been found particularly useful is that of a valve which appears as a capacitance across the tuned circuit, thus retuning the oscillator to the synchronized frequency. The synchronizing effect is then due to the magnitude of this retuning capacitance varying with the phase angle between the synchronizing and synchronized oscillators. The effect has been studied experimentally in this form by connecting a valve showing " Miller effect " across an oscillator tuned circuit. The frequency of the oscillator can then be varied by adjusting the bias on this valve. By use of a rectifier this bias can be derived from the sum of the injected and local harmonics, and it will then depend upon their phase angle. Such a system has been observed synchronized, and in the process of synchronization, on the oscillograph.

The full mathematical theory (in general form), even for simple circuits, is very involved, van der Pol's approximations being the simplest method of approach. Conditions for the existence of a steady synchronized state (without reference to its stability; or transitional and unsynchronized forms) are in this paper approximated to by a more direct method. Suppose the circuit to be broken at some point, say at the grid of a valve, and assume the voltage across this grid to be given by the series

$$A_0 + A_1 \cos \omega t + B_1 \sin \omega t + \ldots$$

The voltages and currents at all other points in the circuit are then obtained in terms of A_0, A_1, etc., also as Fourier series in ω, the synchronized frequency. Finally, the voltage across the circuit which is actually connected to the grid is obtained as:—

$$\alpha_0 + \alpha_1 \cos \omega t + \beta_1 \sin \omega t + \ldots$$

where α_0, α_1, etc., are functions of A_0, A_1, etc., the circuit and valve constants, and the magnitude and phase of the " locking " input energy. Equating all the terms of the two series, a set of relations:—

$$\begin{aligned} A_0 &= \alpha_0, \\ A_1 &= \alpha_1, \\ B_1 &= \beta_1, \\ A_2 &= \alpha_2, \text{ etc.,} \end{aligned}$$

is obtained, and this must hold in practice when the circuit is connected to the grid. Where tuned circuits enable the Fourier series to be approximated to by one or two terms, this " loop " calculation is useful; also, it is exactly applicable to the question of stability at the origin when only the first terms are involved (Nyquist's criteria). In other cases, it is only of rather vague and general application, as the numerical solution of an infinite set of equations with an equal number of unknowns is not exactly practical. It may be noted, however, that whereas a " free " system has frequency ω but no phase angle ϕ_1 to be determined, a synchronized system having no unknown frequency ω will always have a determinate value for ϕ_1. That is, the phase of the synchronized oscillation relative to the synchronizing oscillation will be fixed and dependent upon the relation of synchronizing to free oscillation frequencies.

APPENDIX 2

Suppose the valve anode current is:—

$$i_a = av_g + bv_g^2$$

With sinoidal drive, $v \sin \omega t$,

$$\begin{aligned} i_a &= av \sin \omega t + bv^2 \sin^2 \omega t \\ &= av \sin \omega t + bv^2 \tfrac{1}{2}[1 - \cos 2\omega t] \\ &= \frac{bv^2}{2} + av \sin \omega t - \frac{bv^2}{2} \cos 2\omega t \end{aligned}$$

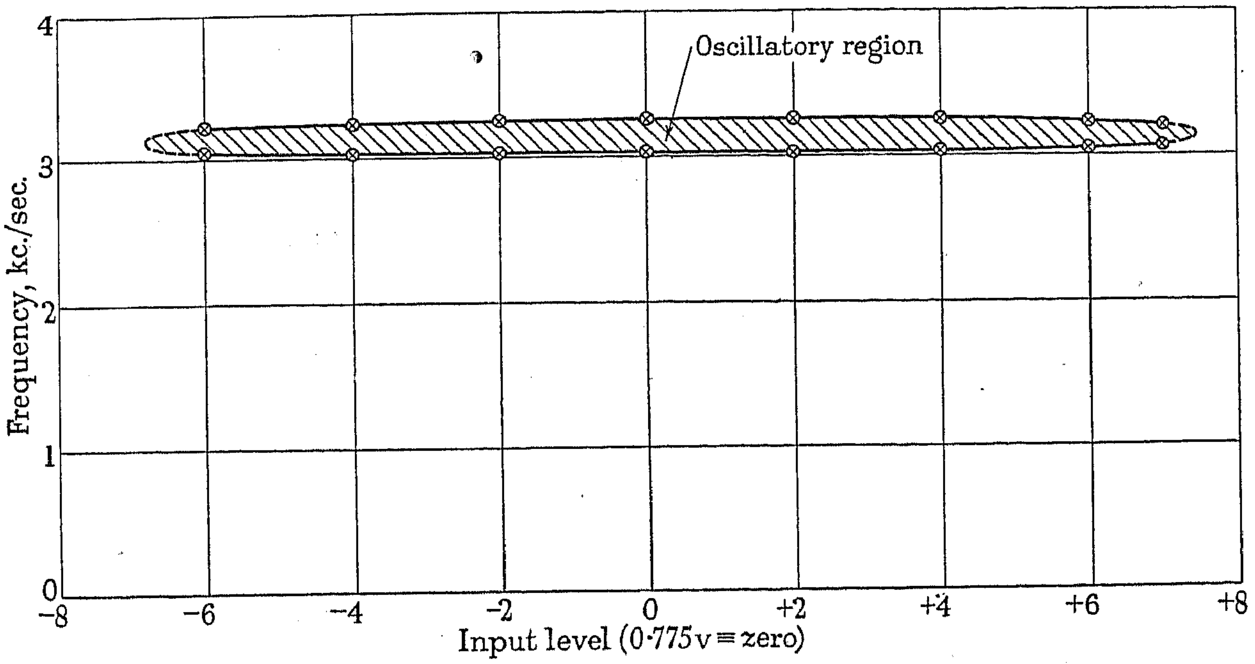


Fig. 4.—Experimental quasi-stable divider; test results.

Anode Currents at 150 V:—
With input $V_1 = 1{\cdot}8$ mA; $V_2 = 1{\cdot}2$ mA.
Without input $V_1 = 1{\cdot}2$ mA; $V_2 = 0{\cdot}3$ mA.

Frequency and level limits:—
150 V (H.T.) ± 3 % on 3 150 c./sec.; $\pm\frac{7}{8}$db. on zero level.
165 V (H.T.) ± 3 % on 3 150 c./sec.; ± 8db. on zero level.
135 V (H.T.) ± 2·8 % on 3 150 c./sec.; $\pm\frac{6}{8}$db. on zero level.

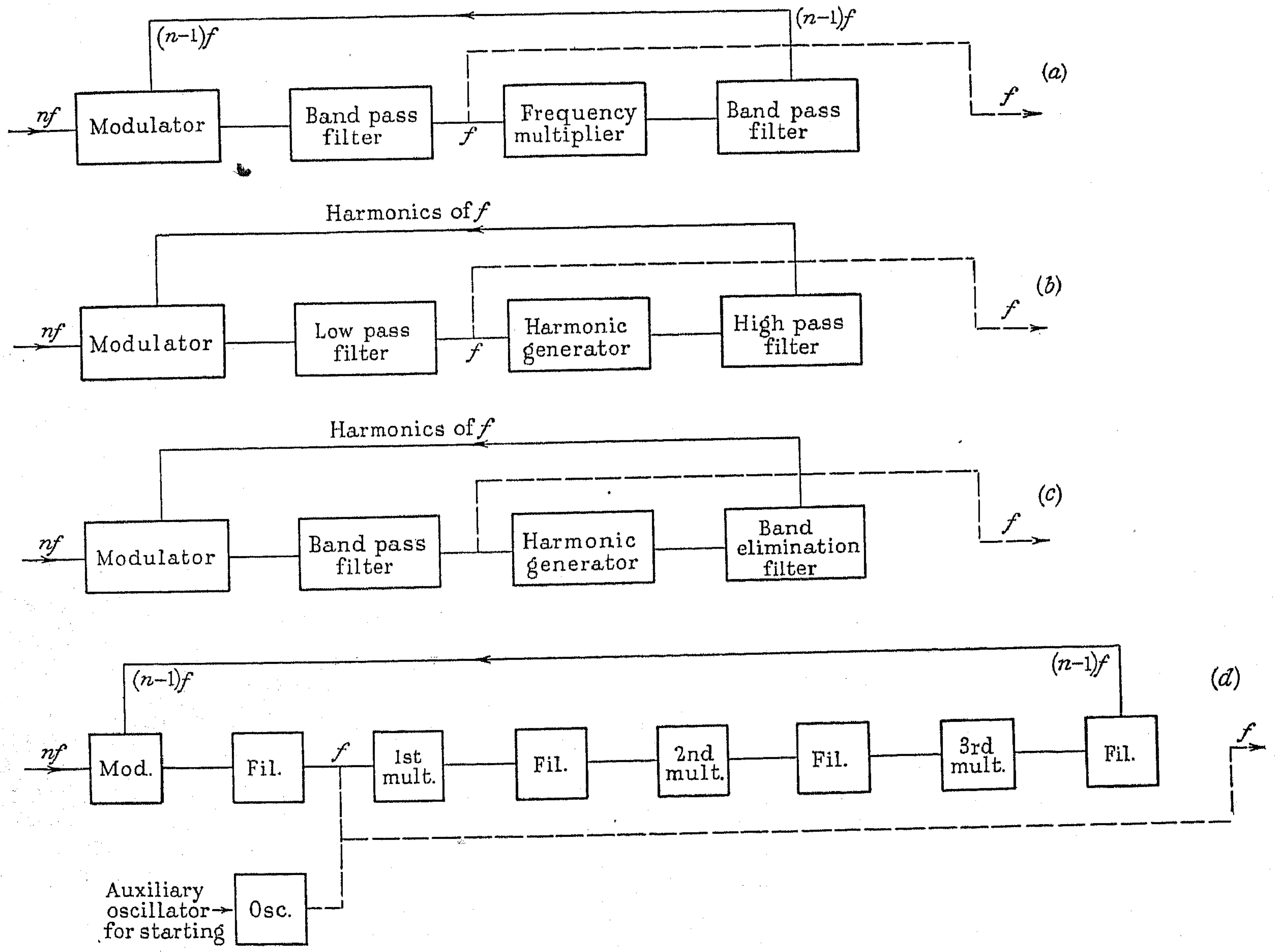


Fig. 5.—Possible developments from simple quasi-stable circuit.

The fundamental term fed back is $(av\sin\omega t)$.

Adding 2nd harmonic $v_g = v[\sin\omega t + k\sin(2\omega t+\phi)]$

$$\begin{aligned} i_a &= av[\sin\omega t + k\sin(2\omega t+\phi)] + bv^2[\sin^2\omega t \\ &\quad + k^2\sin^2(2\omega t+\phi) + 2k\sin\omega t\sin(2\omega t+\phi)] \\ &= av[\sin\omega t + k\sin(2\omega t+\phi)] + bv^2\{\tfrac{1}{2}(1-\cos 2\omega t) \\ &\quad + \frac{k^2}{2}[1-\cos(4\omega t+2\phi)] + k[\cos(\omega t+\phi) \\ &\qquad - \cos(3\omega t+\phi)]\} \\ &= \frac{bv^2}{2}[1+k^2] + av\sin\omega t + bv^2k\cos(\omega t+\phi) \\ &\quad + avk\sin(2\omega t+\phi) - \frac{bv^2}{2}\cos 2\omega t \\ &\quad - kbv^2\cos(3\omega t+\phi) - \frac{k^2bv^2}{2}\cos(4\omega t+2\phi) \end{aligned}$$

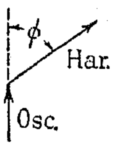
It will be seen that the fundamental term is now $av\sin\omega t + bv^2k\cos(\omega t+\phi)$, the second term being introduced in virtue of the modulation product of $\sin\omega t\sin(2\omega t+\phi)$. Its phase depends upon the angle ϕ of the injected harmonic compared with the fundamental.

APPENDIX 3

It is shown in the analysis that an oscillator of free frequency ω_f may be able to maintain a steady oscillation at a different frequency, ω, if a harmonic of that frequency, $n\omega$, is injected. That it *will* do so, and continue, even if the harmonic frequency changes slightly, is a much more complex question. Suppose the input frequency changes to $n\omega(1+\delta)$. It may be considered that the injected harmonic is still at the exact multiple, $n\omega$, but has changing phase angle, $n\omega\delta t$. As the phase of injected harmonic changes, however, the phase of the fundamental fed back changes, for the intermodulation feed-back depends upon harmonic phase relative to fundamental. As this change of phase of feed-back occurs, the frequency of the oscillator must change, and in a stable synchronized system this change of oscillator frequency to $\omega(1+\delta')$ is such that the phase angle of the harmonic, which originally caused the change, is reduced. As this happens, the frequency changes back towards ω again; and a hunting, which must in practical cases be aperiodic, will occur until the system settles down with the fundamental and harmonics at some new angle giving just the correct sub-multiple frequency, $\omega(1+\delta)$. The process can be described in terms of vectors:—

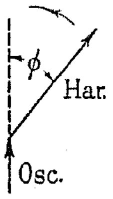
(1) Oscillation (multiplied by n) and injected harmonic; steady state.

Angle ϕ not changing, since system is synchronized.

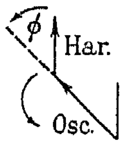


(2) Injected harmonic frequency increased, harmonic vector rotating and ϕ decreasing.

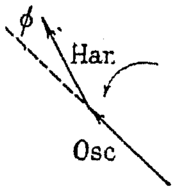
Decreased ϕ, however, increasing oscillator frequency.



(3) Oscillator frequency increased, "oscillation" vector therefore also rotating and "chasing" harmonic; rate of increase of ϕ falling.



(4) New steady state, reduced ϕ giving change of oscillation frequency equal to change of harmonic frequency, and whole diagram rotating with fixed angle between vectors. Rate of rotation equal to frequency change; rate of change of ϕ zero.



Practical synchronization requires, therefore, besides the "existence" condition given, an additional "stability" condition that, when small frequency-changes are made, the oscillator will follow them as described above. The process has been observed, with a special oscillator, on the oscillograph (see Appendix 1).

FLUCTUATION NOISE IN PARTIALLY SATURATED DIODES[*]

By D. A. BELL, B.A., B.Sc., Graduate.

(*Paper first received* 26*th May, and in revised form* 18*th November,* 1938.)

SUMMARY

A previous paper[†] discussed the nature of fluctuation noise in thermionic valves which are either perfectly temperature-limited or perfectly space-charge-limited. It was suggested that the noise should be represented by a " shot noise " formula in the former, and a " thermal noise " formula in the latter. But since perfection is difficult to achieve, the case of more practical value is the valve which is mainly space-charge-limited, but not entirely free from temperature limitation. A method of quantitative estimation of the noise in diodes under such conditions (in terms of the current/voltage characteristic) is the subject of this further paper.

The " ideal " diode would have a characteristic such as that sketched in Fig. 1, curve (*a*), following a $\frac{3}{2}$-power law from the origin to a point $V_a = V_s$, and then changing abruptly to constant current. In an actual valve, however, the curve is more likely to be as shown by the dotted line (*b*), where the current is not zero at $V_a = 0$ (owing to electron emission velocity). The slope of the curve is always less than corresponds to the $\frac{3}{2}$-power law, and the change to temperature-limitation (constant-current condition) is comparatively gradual. From the previous paper we have hypotheses which would permit us to predict the noise from the i_a/V_a characteristic for both parts of curve (*a*); namely, as thermal noise in a certain equivalent resistance at a defined effective temperature for the $\frac{3}{2}$-power law (space-charge-limited) region, and as simple shot noise for the constant-current (temperature-limited) region. But we have to face the problem of curve (*b*), which does not belong wholly to either of these regions.

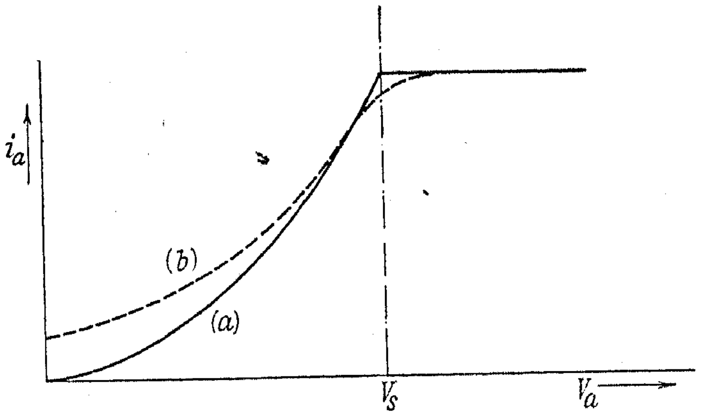


Fig. 1

Consider now a hypothetical system of two diodes of identical construction but different cathode temperatures, one of which, (*a*), is perfectly space-charge-limited, while the other, (*b*), is perfectly temperature-limited throughout the range of anode voltage chosen for examination. The temperature-limitation of (*b*) is suggested in Fig. 2 by the resistance in series with its heater, giving a lower cathode temperature and less emission. The currents in the two valves are accordingly given by the expressions

$$i_1 = aV_a^{3/2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

$$i_2 = b \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

and the total current i_0 will be

$$i_0 = i_1 + i_2 = aV_a^{3/2} + b \quad . \quad . \quad . \quad (3)$$

Now suppose it is impossible to measure separately the currents i_1 and i_2; it will still be possible to determine them from a series of observations of i_0 at different values of V_a, by making use of the slope of the i_0/V_a characteristic. Thus

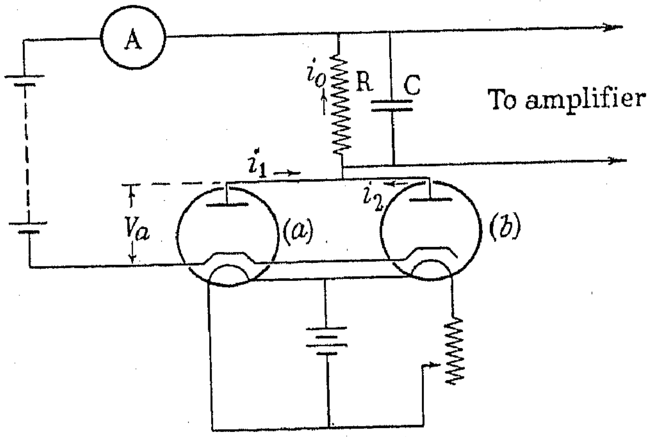


Fig. 2

$$\frac{di_0}{dV_a} = (3/2)aV_a^{\frac{1}{2}} = \frac{3i_1}{2V_a}$$

$$\therefore \; i_1 = \frac{2V_a}{3} \cdot \frac{di_0}{dV_a} = \frac{2V_a}{3R_a} \quad . \quad . \quad . \quad (4)$$

In fact the measurable slope conductance di_0/dV_a of the combination is equal to that of valve (*a*) alone, since valve (*b*) has zero slope conductance. By simple subtraction we then write

$$i_2 = i_0 - i_1 = i_0 - \frac{2V_a}{3R_a} \quad . \quad . \quad . \quad (5)$$

so that both constituent currents are known. The value of this is seen when we replace the two separate idealized valves by a single imperfect diode, having, for example, a non-uniform cathode temperature, in which some parts of the cathode reach temperature-limitation while other parts are still space-charge-limited; by means of equations (4) and (5) we can determine what proportions of the anode current of such a diode are to be attributed to each regime.

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[*] Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 723.

[†] " A Theory of Fluctuation Noise," *Journal I.E.E.*, 1938, vol. 82, p. 522, and *Proceedings of the Wireless Section*, 1938, vol. 13, p. 97.

Before investigating the fluctuation noise to be expected from the system, we must decide what units are to be employed in specifying it. The obvious course of using absolute units is open to the objection that it does not give a ready comparison between valves of different dimensions [different values of the constant a in equation (1)]. Following the example of previous authors, therefore, the results are preferably expressed as the ratio of the observed noise to the noise that would be generated by an equal current i_0 flowing through a temperature-limited valve connected in the same measuring circuit. This ratio has sometimes been called the "smoothing factor"; but that name is undesirable since it is suggestive of the "thermal noise plus shot noise" theory, which is not consistent with the hypotheses adopted by the author. The ratio in question will therefore be termed simply "the noise ratio." It is now necessary to distinguish between two possible noise ratios. First, one may take the ratio of the mean-square values, and this is the factor which Dr. F. C. Williams denotes by A;* alternatively, one may take the ratio of the root-mean-square values, and this is the factor denoted by F in the work of Prof. W. Schottky and his colleagues.†

The mean-square value is obviously the factor of greater primary significance, but the derived root-mean-square value will have practical advantages (as pointed out by Prof. Schottky) when the theory can be applied to amplifying valves instead of diodes. For the noise

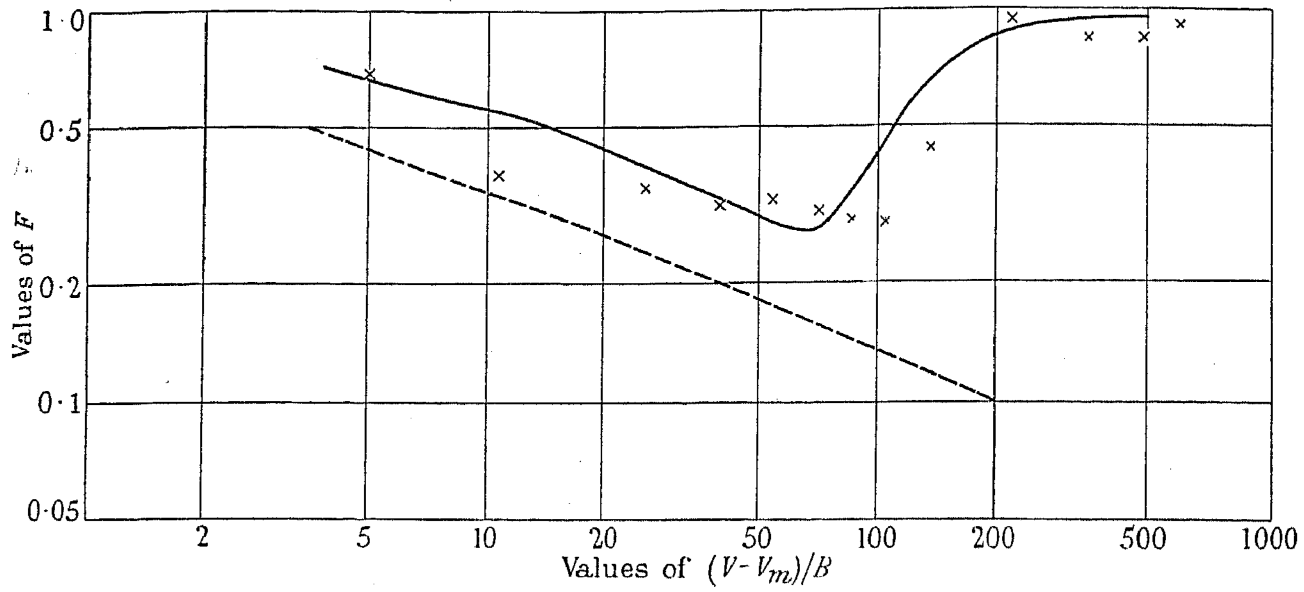

Fig. 3.—" Rauschröhre 6."

Calculated curve ————
Jacoby and Kirchgessner measured points × × ×
Schottky calculated curve of lower limit of noise - - - - - -

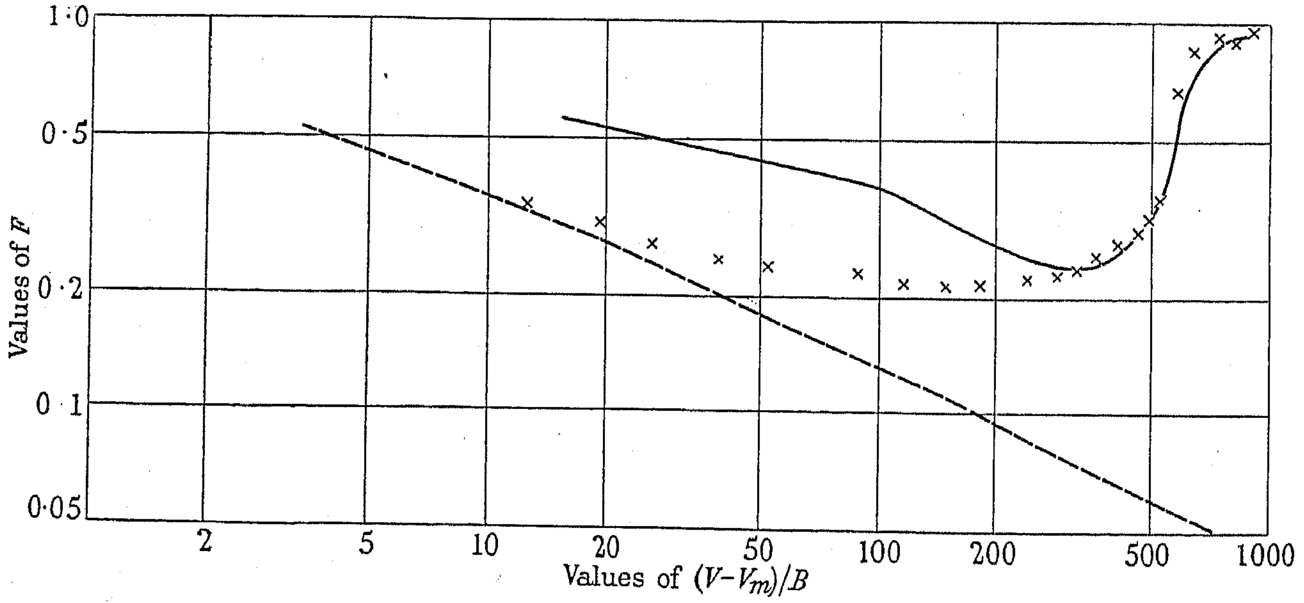

Fig. 4.—"Rauschröhre 7."
(Details as Fig. 3.)

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* See, for example, *Journal I.E.E.*, 1936, vol. 79, p. 349, and *Proceedings of the Wireless Section*, 1936, vol. 11, p. 316.

† See, for example, *Wissenschaftliche Veröffentlichungen aus den Siemens-Werken*, vol. 16, part 2.

problem normally arises in voltage amplifiers, whose signal output is specified linearly in volts, not as a squared value, so that signal/noise ratios will be more easily calculated if r.m.s. noise values are used. In practice F is obtained when required simply by taking the square root of A.

Returning to Fig. 2, let us calculate the total fluctuation current to be expected from the two currents i_1 and i_2, which are assumed both to flow through a noise-measuring system consisting of the load circuit R, C, and an amplifier. Considering a narrow frequency band df, valve (b) will contribute a " shot noise " fluctuation current of mean-square value

$$\overline{I_s^2} = 2i_2e\,.\,df \quad . \quad . \quad . \quad . \quad . \quad (6)$$

Now it was shown in the previous paper[*] that a valve of $\frac{3}{2}$-power law and slope resistance R_a appears as a thermal-noise source of mean-square e.m.f.

$$\overline{V^2} = 4(1\cdot2)R_ak(T/2)\,.\,df \quad . \quad . \quad . \quad (7)$$

The mean square fluctuation *current* in such a valve must therefore be

$$\overline{I_t^2} = \frac{\overline{V^2}}{R_a^2} = \frac{4(1\cdot2)kT}{2R_a}\cdot df \quad . \quad . \quad . \quad (8)$$

Since we are concerned with mean-square values of random currents, the total fluctuation noise is the sum of the components from valve (a) and valve (b), namely

$$\overline{I_0^2} = \left\{2i_2e + \frac{4(1\cdot2)kT}{2R_a}\right\}df \quad . \quad . \quad . \quad (9)$$

But the noise from the whole current i_0 if it were entirely temperature-limited would be

$$\overline{I_x^2} = 2i_0e\,.\,df \quad . \quad . \quad . \quad . \quad . \quad (10)$$

On dividing (9) by (10), the mean-square noise ratio is

$$\frac{\overline{I_0^2}}{\overline{I_x^2}} = A = \frac{i_2}{i_0} + \frac{1\cdot2kT}{i_0eR_a} \quad . \quad . \quad . \quad . \quad (11)$$

The factor F is then obtained as the square root of A.

This method of calculation has been applied to the two diodes described as " Rauschröhre 6 " and " Rauschröhre 7 " in a paper by H. Jacoby and L. Kirchgessner,[†] and Figs. 3 and 4 show a comparison of their published measurements of noise ratio with the values of F calculated by the method described above from their published i_a/V_a characteristics. In view of the fact that the published characteristics are printed to a size of graph of $2\frac{3}{8}'' \times 3''$ only, and as it was necessary to re-plot these and apply graphical differentiation to obtain values of R_a,[‡] the agreement of theory with experiment is considered to be good. The abscissae are as in Jacoby and Kirchgessner's paper, in units of $(V - V_m)/B$, where $(V - V_m)$ is the potential difference between anode and potential minimum, and B is the equivalent in volts of the mean emission energy of the electrons ($B = kT/e$). The potential minimum has been calculated in terms of the ratio of the observed anode current at any voltage to the total cathode emission, the latter being assumed to be equal to the saturation value of anode current reached at high anode voltage.

As an indication of the values involved, a set of figures corresponding to the curve in Fig. 3 is given in the Table.

The best basis of comparison between diodes and triodes is in terms of the ratio of working anode current to total cathode emission. Let us assume that a triode amplifying valve has a working range of anode current from 10 % to 50 % of cathode emission. Then in Fig. 3 and the Table it will be seen that over this range for the diode the temperature-limited component of current i_2 contributes 60 % to 90 % of the noise, though it is only of the order of 10 % to 15 % of the anode current. This indicates at once the liability of any commercial valve to have a noise level far above that predicted by the thermal theory on the assumption of complete space-charge limitation. It also indicates the difficulty of checking the theory by published results of other workers, since very precise knowledge of the i_a/V_a characteristic is necessary to determine i_2 with the required accuracy. The d.c. characteristics of the valve have not hitherto been considered of fundamental importance to the theory of fluctuation noise, and the author has so far been unable to find published results for triodes which give suitable data.

Table

V		R_a	i_1	2	$\frac{6kT}{5eR_a}$ $\times 10^{-3}$	F
1·35	0·192	7 500	0·12	0·072	0·028	0·722
2·60	0·391	5 500	0·316	0·075	0·038	0·538
3·77	0·637	4 550	0·553	0·084	0·046	0·452
6·25	1·23	3 650	1·14	0·09	0·057	0·346
11·09	2·58	3 000	2·46	0·12	0·069	0·272
13·4	3·40	2 940	3·03	0·37	0·071	0·36
13·9	3·5	2 940	3·04	0·46	0·07	0·388
17·0	4·57	2 850	3·98	0·59	0·073	0·38
20·5	5·66	3 300	4·15	1·51	0·063	0·526
34·5	7·41	10 050	2·29	5·12	0·02	0·841
51·7	8·20	29 500	1·17	7·03	0·007	0·926
70·5	8·71	42 000	1·12	7·59	0·005	0·934
89·7	8·89	57 000	1·05	7·84	0·004	0·946

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[*] *Loc. cit.*

[†] *Wissenschaftliche Veröffentlichungen aus den Siemens-Werken*, vol. 16, part 2, p. 42.

[‡] A curve of R_a against i_a is plotted in the published paper, but it was apparently obtained by measurement on a bridge; although satisfactory for the straighter parts of the characteristic, it has not the high precision required for the present purpose in the region of the upper bend. The graphical method therefore had to be used to obtain values of R_a.

RADIO-TELEGRAPHY AND RADIO-TELEPHONY*

By C. E. RICKARD, O.B.E., Member.†

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(1) FOREWORD

In compiling a review of progress of radio-telegraphy and radio-telephony during the past few years, which have been notable for steady advance rather than outstanding invention, one is faced with two main problems: (1) where to start and where to finish, in relating such progress as should be recorded; and (2) what relative importance should be given to the numerous branches and ramifications which go to make up the science, art, and practice, which form the subject matter of such a review.

Progress in the science of radio may be held to cover advance of knowledge in theory and technique, but progress in the art, while dependent upon progress in the science, also depends for its practice to a large extent upon the commercial application of such improvements as may have been made in the past. Further, the commercial aspect of progress in the art of radio, though mainly controlled by economics, is also governed by the application of international regulations in the practice of radio communication.

As we are concerned mainly with the art of radio-telegraphy and radio-telephony, the author may perhaps be forgiven if he selects international regulations as the starting-point of his discussion, since progress in this direction enforces, by mutual agreement between the nations, improvement both in the art and in the practice of radio communication; and, seeing that a new international agreement has recently been reached at Cairo, the occasion to discuss the effects of the 1938 Convention is opportune.

(2) INTERNATIONAL REGULATIONS

The recent plenary International Conference to review the Telecommunications Convention and General Regulations made at Madrid in 1932 was held at Cairo in the spring of 1938. The Conference commenced on the 1st February, and was opened by H.M. King Farouk at the Heliopolis Palace Hotel.

The work of the Conference entailed much study of the business aspects of telecommunication, including such matters as telegraph and telephone tariffs, communication procedure, traffic routing, and many other important factors in the art and practice of telecommunication in all its branches—problems with which we are not concerned in this review. It also included, however, a very important item, namely the revision of the Madrid General Radio Communications Regulations in the light of present-day needs resulting from progress made during the past 5 years, such progress having led to increasing congestion in the ether owing to the tremendous expansion of radio services during this period.

Bearing in mind the importance of the situation thus described and, in particular:—

(*a*) the growth of aerial transport, of which radiocommunication in all its forms is a most important ancillary service;

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* A review of progress.
Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 368.
† Marconi's Wireless Telegraph Company, Ltd.

(*b*) the ever-insistent demands of broadcasting, which have gathered added strength from the increasing weight of public opinion and political activity;

(*c*) the further development of the marine services in all branches, but especially in the short-wave trawler telephony and radiogoniometric services;

the actual modifications made at Cairo to the Madrid General Radio Communication Regulations were considerably smaller than expected.

Nevertheless, considerable changes have been made:—

(i) By virtue of the modifications made in the allocation of frequencies to the various services.

(ii) By the tightening-up of the regulations in respect of the frequency tolerances attributable to the various types of transmitter corresponding to the different services they perform.

The first real allocation of frequencies may be said to have been made at Washington in 1927, largely as a result of experience gained during the Great War and subsequent years, and as a tentative effort to distribute the available frequencies amongst the claimants whose services had comparatively recently (at that time) begun, namely radiogoniometry, broadcasting, radiotelephony for small ships, short-wave point-to-point services, and amateur activities.

The frequency allocations thus adopted were largely influenced by the positions in the radio-frequency spectrum already taken up by the said new services.

At Madrid the frequency allocations were slightly modified to meet growing needs, but frequencies still remained allocated in comparatively wide bands to be shared among various classes of services. This was particularly so in the case of the mobile bands whose frequencies were shared by marine and aircraft services, irrespective of the fact that the types of apparatus practicable for these very different vessels vary considerably in size, weight, and consequently in their technical characteristics.

Great advance was therefore made at Cairo in regard to re-allocations, exclusive bands of frequencies being assigned, in particular to aircraft. Moreover, an effort has been made to facilitate further the solution of the problem of the expansion of the aeronautical services by dedicating exclusive and specific bands of frequencies to the various aircraft routes in contemplation.

Thus the practical result of the revision of the Madrid allocations, so far as aeronautical and marine services are concerned, is that the aircraft radio services have been given the exclusive use of one part of the former mobile service bands and the marine services the exclusive use of the remainder. This separation of marine and aeronautical services, together with the adoption at Cairo of definite tolerance regulations for mobile transmitters, should tend to minimize interference between the two service channels.

Concerning the marine services, some of the changes in the regulations were of necessity subject to commercial considerations. A considerable tightening-up of regulations had been forecast before the meeting, but during the past years a world slump has been experienced and, since the cost of new equipment for what would have been the bulk of the world's shipping would ultimately fall on shipping interests all over the world, too drastic changes could not be made, for economic reasons.

Nevertheless, definite tolerance limits for marine transmitters have now been adopted by the Cairo Conference, whereas at Madrid the recommendations inserted in the General Radio Regulations served only as a guide and not as a definite regulation.

The application of a regulation fixing the tolerance of a ship station entitled to transmit on any frequency within the mobile band is not feasible. The difficulty was overcome at Cairo by a special definition of tolerance as applied to marine transmitters, wherein it is laid down as "the limit of variation from the starting frequency during a 10-minute transmission."

Although this may not appear very drastic, it is to be remembered that most of the frequency drift of the normal simple transmitter occurs during the first 10 minutes, and it is thought that the general effect of this definition, coupled with a new regulation stipulating for the equipment of ship installations with a good wavemeter or its equivalent, cannot fail to be advantageous, while at the same time the employment of an economical form of transmitter still remains possible for ship use.

Ships working in the shared bands (fixed and mobile services) must observe the fixed station tolerances, which impose much higher technical requirements on the transmitter. Progress in this direction is still comparatively slow, since high-class apparatus of the order required is only commercially practicable on first-class passenger vessels, where the extent of the telegraph and telephone traffic to and from such ships justifies the considerable capital expenditure involved.

Technical opinion is, perhaps naturally, biased in favour of the abolition of spark (Type B) transmission for all purposes except for distress signalling, but the bulk of the world's shipping, so far as number of vessels is concerned, consists of cargo vessels compulsorily fitted under Safety of Life at Sea Regulations, whose owners demand the right to use their radio transmitters for the rare messages they may wish to send. Such an argument is difficult to resist, and thus a compromise has been found by limiting the transmission of spark to sets below 300 watts input from the alternator terminals and to operation on 375 kc./sec. (800 m.) for radiogoniometry, 425 kc/sec. (706 m.) for traffic, and 500 kc./sec. (600 m.) for calling and distress signalling.

Progress is thus being realized by the gradual supersession of high-power spark transmitters, such as has taken place during 1938 and must be accelerated and completed by the end of 1939, by which time, so far as British shipping is concerned, at least 1 000 high-power spark sets will have disappeared.

Consideration was also given at Cairo to the large number of small ships operating radio-telephone and radiotelegraph equipments in the region below 1 500 kc./sec. (200 m.). An international calling and distress frequency of 1 650 kc./sec. (181·8 m.) was assigned for such services, and administrations have undertaken to organize a watch to be kept on this frequency at suitable coast stations. This should enhance the "little ship" services and foster further development.

But for a small number of narrow bands relinquished to broadcasting, no marked change was made in the frequency allocations for the fixed services. Although the growth of these services has continued in recent years, yet technical progress in improved methods of frequency stabilization, combined with careful organization of individual allocations, has fortunately rendered the provision of more room unnecessary. The only important change made was the adoption of much more stringent frequency-tolerance regulations for fixed stations.

In the search for new channels, free from congestion and parasitic interference, use is now being made of the waves below 10 m., amongst which the broadcasting of television and the inauguration of inter-island telephony circuits by the British Post Office on waves ranging from 2 to 7 m. or thereabouts are notable examples.

A further sign of progress indicated by the new allocation of frequencies at Cairo is the assignment of all frequencies up to 200 Mc./sec. (1·5 m.) to broadcasting, television, radio balloon sounding, fixed, aeronautical, marine, and amateur services, whereas the Madrid allocations were limited to frequencies up to 60 Mc./sec. (5 m.).

(3) MARINE SERVICES

As has been mentioned, the development of improved ship installations is very closely bound up with the activity of shipping. In consequence, there has been a slowing-up in the application of new designs and improved marine equipments, with the notable exception of expansion in short-wave fittings for world-wide communications and the development of radio-telephone equipment for trawlers and coasting steamers.

Improved Stability

One of the principal difficulties in short-wave working under ship conditions has been the problem of accurate tuning of the transmitter to the degree of precision required by international regulations and also to ensure that the signal is quickly picked up by the coast station without causing interference with other services. By making use of the fact that the marine short-wave bands are arranged in harmonic relationship, it has proved possible to develop a simple monitoring scheme which permits of immediate and direct checking of the tune of the transmitter by reference to a single primary quartz-crystal standard of 345 kc./sec. This relatively low-frequency crystal is of the low-temperature-coefficient type and affords a very satisfactory answer to the question.

Coast Stations

The considerable increase in the number of ships equipped with short-wave radio-telegraph equipment has necessitated corresponding increases in coast-station facilities, and as a result the world-wide short-wave telegraph service operated with ships from the stations at Portishead and Burnham has grown steadily, together with the installation of further transmitting and receiving equipments. The latest transmitters installed are capable of emitting twice the carrier power of the earlier equipments, can each transmit two messages simultaneously, and are capable of very rapid wave-changes such as are necessary in dealing with traffic from ships located in all parts of the globe. The latest transmitter is crystal-controlled by means of quartz crystals of almost zero temperature-coefficient operating at the fundamental radio frequencies assigned to the transmitter, thereby enabling the usual harmonic-generating stages to be dispensed with. Further economy in space is effected by employing pentode power valves in the preliminary amplifying stages.

A simple "quick search" type of short-wave receiver has been developed to speed up the clearance of traffic on this service, the broad principle of the receiver being to convert the incoming frequencies by wide-band demodulators into frequencies lying in the medium-wave broadcasting band and then to employ a more or less conventional type of broadcasting receiver.

Work has been going on steadily with the modernization of the shorter-range coast-station equipments to enable these to comply with the latest permissible frequency tolerances. Some stations have recently been equipped with transmitters having all frequencies stabilized by quartz crystals and designed for operation on both telegraphy and telephony.

Telephony

The increased use of radio-telephony in small vessels such as coasters and trawlers has necessitated much additional organization. Only limited bands are available for this service, into which the ship-to-shore and inter-ship communications of all nations must be packed. The difficulty can be realized when it is understood that there are more than 2 000 vessels of this kind fitted with telephone installations, and nearly all operate in congested and narrow seas. Much has been done through the operation of the North Sea and Baltic Agreements, which allocate frequency bands on a regional basis, thereby reducing the mutual interference. These arrangements are being followed up by the design of new transmitters better able to control their radiated frequencies in accordance with requirements.

Trawlers and Small Ships

Owing to the normal growth of traffic and to the increasing use of radio communication by trawlers and small craft it has been necessary to provide means whereby coast stations may send and receive simultaneously at the same station. This has been done by erecting the receiving aerials about 1 mile from the transmitting station; by connecting these receiving aerials to the receivers by means of low-loss coaxial cables; and by providing selective receivers in a screened room. As a result of these precautions interference from the local transmitters, even where these are of high power, has been reduced to a negligible value. This system, which has been installed at the Wick station, has proved to be entirely satisfactory, and arrangements have been put in hand for similar systems to be installed at other coast stations. Arrangements have also been made for the direction-finding aerials to be remotely situated from the receiving stations at the Lands End and Wick stations, because by this means the arc over which reliable bearings are obtainable can be considerably increased.

A continuous distress watch for trawlers on 181·8 m. has been instituted at all coast stations, and a device has been developed which causes the receiver to search continuously over a band of about 30 kc./sec. every half-minute to cater for ships which may accidentally be transmitting off their nominal frequency.

Express Steamer Equipment

The addition of the " Queen Mary " to the mercantile marine has been a milestone from the wireless engineer's point of view, in that a most complete and effective installation has been fitted. The " Queen Mary " equipment is designed to meet the special requirements of vessels of her class, and will serve as a model for installations of this type. It has recently been very fully described in the paper by Commander Loring and Messrs. McPherson and McAllister.* The installation design was reached after a most careful and scientific survey of the needs of an express steamer such as the " Queen Mary." The manner in which these needs have been met with present-day technical knowledge, having due regard to the special conditions on board ship, is detailed in the paper.

The equipment consists of: (1) Short-wave telegraph and telephone transmitter. (2) Medium-wave telegraph transmitter. (3) Long-wave telegraph transmitter. (4) Four operating positions, with receivers for providing adequate facilities for reception on all wave-bands. (It is possible to key any transmitter on any selected spot wave from each position.) (5) Telephone terminal and privacy equipment. (6) Position for automatic high-speed telegraphy reception and transmission. (7) Emergency transmitter. (8) Auto-alarm. (9) Direction-finder. (10) Broadcast receiver.

As the transmitting room is some 400 ft. distant from the receiving and operating room, remote control of the transmitters is provided. The operators are by this means able to select any of the spot waves required, by a simple dialling operation in the receiving cabin.

Interference with Radio Working on Board Ship

With the increasing use of music reproduction and short-wave reception on board ship, the suppression of " ship noises " (or, in other words, electrical interference with radio working) has become a problem of major importance. Such interference is caused by electromagnetic induction between ships' cables and wiring, usually as a result of the minute sparks which occur at the brushes of all rotating electrical machinery or at the contacts of switches and circuit-breakers of all kinds. The problem has been given careful consideration by the various specialized committees concerned with the revision of the I.E.E. Regulations for the Electrical Equipment of Ships.

(4) DIRECTION-FINDING SERVICES

The rapid growth of both military and civil aviation has greatly stimulated progress in direction-finding. Wireless navigation is carried out by observation of the mobile transmitter from a ground direction-finding station, by the observation of a fixed beacon station from a direction-finder installed in the moving vessel, or by the use of some type of equi-signal or " beam " type of ground transmitter to lay down a course for the aircraft or ship. The fact that aircraft need bearings at long ranges both by day and by night and may require the observations to be made in the short-wave band has led to a close study of methods of direction-finding which are free from the errors encountered under these conditions when using the conventional rotating loop or radiogoniometer.

Adcock Direction-finders

Development has continuously taken place as regards the solution of technical difficulties in the application of the Adcock principle. This basic idea depends for its success upon a recognition of the fact that all random direction-finding errors of practical importance are due to reception by portions of the aerial system lying in a horizontal plane. Two main methods of countering the effects are employed. We may either arrange the horizontal members of the receiving aerial so that they are symmetrically disposed in such a way that the voltages induced by the incident electromagnetic waves are always equal and opposite, so that their effects cancel, or we may electrically screen all horizontal conductors from the effects of the arriving radiation. From the engineering standpoint each way has its merits. In effectiveness in removing the errors there is little to choose between them.

The polarization-error-free direction-finder has now been established on a firm foundation; many such equipments have been erected both at home and throughout the Empire, and are playing an important part in aerial navigation.

Mobile Installations

Improvements in mobile direction-finding equipment for ships and aircraft have been mainly in the direction of neater and more convenient designs suitable for use by the navigating officer, supplemented by the study and incorporation of various methods of obtaining visual indication of the bearing of a selected beacon station. The latest types of direction-finder are tuned by means of a single knob, and the beacon stations employed by a ship on any particular run are as easily picked up and identified as the stations on the familiar home broadcast receiver. Visual indication of the bearing of the beacon station, in general, takes the form of some pointer instrument or cathode-ray tube which tells the observer when he has oriented his pointer correctly to read off the bearing required. Several types of direct-reading visual indicator in which the bearing itself may be directly read on a scale are under development, and some equipments of this type have emerged from the laboratory and are undergoing practical trials.

Cathode-ray Methods

In this connection mention must be made of the " comparator " method of using a cathode-ray oscillograph as a direct-reading visual direction-finder. The result is achieved by employing two frame or Adcock aerials in mutually perpendicular planes, each being separately tuned and the received currents being separately amplified before application as independent voltages to the deflecting pairs of plates of the oscillograph. By careful attention to the removal of amplitude and phase distortion in the two chains of receiving circuits it is possible to make the trace on the screen of the tube indicate the bearing of the station under observation, the combination of the effects from the two frames being exactly analogous to that in the familiar Bellini–Tosi radiogoniometer, but having the advantage of giving an instantaneous visual indication which is continuously observable. This method has been very closely studied and is finding considerable application in cases where rapidity and directness of observation outweigh such considerations as cost, simplicity, and wave-range.

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* *Journal I.E.E.*, 1937, vol. 81, p. 183, and *Proceedings of the Wireless Section*, 1937, vol. 12, p. 198.

Equi-signal Beacons

The equi-signal beacon continues to be the subject of study and development both at home and abroad. It finds its main applications in air-route guidance and in aerodrome approach and landing aids, about which more will be said later in this review.

Maritime Radio Beacons

The period under review has seen a steady growth in the number of beacon stations available for marine navigation. This activity has been accompanied by improvements in the performance of the stations, especially in regard to (i) frequency stability and (ii) methods of modulation. These points are now of special importance inasmuch as the increase in the number of beacons is rendering the packing of the stations into the small frequency band available increasingly difficult. A further factor in the move towards accurate carrier-frequency stabilization is the greater personal use by navigating officers of the marine direction-finder. Under these circumstances it is essential that the carrier frequency of any beacon should be accurately controlled in order that it may tune exactly at the point on the tuning scale of the direction-finding receiver at which it is expected to appear. Variation in this respect might lead either to wrong identification of a beacon or to failure to receive it.

The utility of the beacon services has also been enhanced by the provision of two additional facilities for direction-finding calibration. For this purpose a programme of transmissions has been arranged to take place daily, the beacons sending at specified times for 1 hour in the morning and 1 hour in the afternoon, each programme consisting of 10 successive fog transmissions. In addition to this an "on request" service has been commenced, whereby any beacon will transmit the fog programme continuously on request by a ship requiring calibration. After notification of the authority concerned, the ship, on arrival off the beacon, gives a visual and sound signal (black ball at mast head and blast sequence on ship's whistle); the beacon will then commence to transmit and will continue to do so until the ship notifies the beacon that the calibration is complete.

Beacon Equipments

A number of Trinity House beacons have been brought up to date by the use of high-precision master oscillators whose stability is of the order of 1 in 25 000 or better, all new installations being now equipped with quartz oscillators in order to comply with the Cairo requirements. Similarly on British lightships the process of replacing existing telephone sets to meet the new regulations is proceeding. Such sets now have a tolerance of 1 in 10 000.

The original remote-controlled fog gun in the Clyde is also being modernized by the use of a crystal-controlled transmitter, which now permits the use of a receiver so selective that false alarms are practically a thing of the past.

The Northern Lighthouse Board has also been active in increasing the utility of the service, and beacons have been put into service at Barra Head and Port St. Mary (Isle of Man). A further station is in course of erection at the North Uist Lighthouse. The Barra Head station is of interest in that the aerial crosses a gully and is, in consequence, about 700 ft. above sea-level. This station has proved to have a range of over 200 miles.

The same authority has also been actively proceeding with the equipment of lighthouses with radio-telephone equipment to maintain communication with the shore.

(5) AERONAUTICAL COMMUNICATIONS

There has been a steady expansion in the normal aircraft-to-ground communication facilities, in keeping with the growth of the commercial services. It should, however, be remembered that the development of this type of facility must always take second place to meteorological services and navigational and landing aids. In any case the demand for telephonic or telegraphic contact between passengers and ground is small, since the present continental journeys are of short duration. With the development of long-distance flying this situation may change.

Aircraft Direction-Finding

Much progress has been made in direction-finding both from the ground and in the air, and this section of the work is dealt with under "Direction-finding." A directional aid which is wholly specialized in the aircraft field is the system of aerodrome approach and landing guidance in bad visibility, which has been the subject of parallel development in various countries.

Briefly, the principle of the system is to provide, from an ultra-short-wave beacon at a selected spot on the aerodrome, a radiated space pattern at an angle from the ground in the vertical plane, such that a pilot can be guided accurately along a sharply defined path of interlocking signals to the aerodrome, and thereafter, by watching the movements of a second needle indicator, receive some check on his angle and rate of descent to the landing runway. The position of the aircraft relative to the aerodrome is defined by means of marker beacons generally located along the approach path, one at the edge of the landing ground and another a few miles away, their duty being to indicate, by means of ultra-short-wave beams directed vertically, known points at which the pilot should adjust his height preparatory to the descent. Although many satisfactory demonstrations have been given of the utility of the system, it requires more investigation before the landing problem can be regarded as completely solved. Much active work is in progress with a view to improving the reliability of the method, to simplify the equipment in the aircraft, and to make the indications as positive and unambiguous as possible. This work covers the development of interlocking-pattern beacons and the study of observed day-to-day variations in the position of the glide path.

Intercontinental Flying

The recent flying-boat flights across the Atlantic in both directions have been of a much more routine character, and undoubtedly presage the establishment on a commercial footing of intercontinental services involving difficult transits. This work has led to the development of much more ambitious aircraft equipment operating in the short-wave bands for long-distance communication. An example of such an aircraft transmitter is that installed in the Imperial Airways flying boats "Caledonia" and "Cambria." This transmitter covers a waveband of from 15 to 1 100 m. in six ranges, and delivers 60–75 watts to the aerial. Instantaneous operation on one of six pre-selected waves is provided, and the whole transmitter occupies less than 2½ cu. ft., and weighs just over 50 lb. These developments have resulted in corresponding changes in receiver designs to cover much wider wavebands and to give the best possible performance for short-wave long-distance communication. Increasing use is being made of both mechanical and electrical systems of remote control for aircraft apparatus. In planning a commercial aircraft equipment it is advantageous to be able to place the instrument itself (transmitter, receiver, or direction-finder) in the most suitable position from the point of view of the layout of the accommodation, and to lead the controls to the pilot's or navigator's position via easily-fitted flexible connections. In consequence, there have recently been developed various aircraft sets embodying this principle and using either mechanical or electrical means of obtaining the control over the connecting links.

(6) LONG-DISTANCE TELEGRAPH CIRCUITS

Progress in telegraph circuits has been mainly confined to the study of traffic requirements and the development of engineering and organization methods hand-in-hand, to give the best service between various points in the world which economic considerations indicate.

Minimum-delay Circuits

A very good example of a development of this kind is afforded by the so-called "Shorter's Court" telegraph circuit linking the London and New York Stock Exchanges. In this case the object is to secure the interchange of information with the absolute minimum of delay which can be achieved. For the purpose of this circuit special telegraph offices have been established adjacent to the New York and London stock exchanges. In London, the office is in Shorter's Court, and is directly accessible to the Stock Exchange itself. Members hand their messages to the transmitting clerk, and arrangements are such that a circuit is always available for instant use.

For the service it has been found necessary to reserve five separate in and out circuits between London and New York. These wireless circuits are kept permanently established across the Atlantic and as many channels as desired are capable of being operated simultaneously electrically in parallel, as circumstances require. This parallel operation is adopted to secure the maximum freedom from questions (RQ's) and replies (BQ's) relating to transmission and reception faults. The engineering precautions taken to avoid interruption of the service are very complete, and include such factors as completely independent power supplies for the circuits, and different wavelengths for the radio links, while separate landlines connect each individual radio transmitter and receiver to the Stock Exchange Office.

Various points affecting the normal design of terminal apparatus have received attention. For example, a special tape transmitter is employed which operates with a delay of only 10 centre holes between the sending point and the keyboard perforator. This development avoids the delay in the storage loop which normally exists between the perforator and the transmitter. Also, at the receiving position the actual tip of the writing siphon is directly visible, at a readable position, to the receiving typist, and so no delay in transcription occurs.

For timing, special time stamps which endorse the messages with times of handing-in and of reception to an accuracy of ± 6 sec. are employed.

Since the offices are not actually in the Exchanges, but adjacent, messages have to be carried by hand to and fro. To speed up this work the messenger boys are fitted with running shoes, and in and out traffic is separated by a white line on the connecting staircase and passage to avoid delay due to collisions.

As a result of all these developments it has proved possible for a member of the London Stock Exchange to hand in his message for a member of the New York Stock Exchange and to be handed a reply in 24 sec. The average delay is less than 60 sec.

Printing Telegraphy

The consolidation of the cable and wireless circuits has led to the adaptation of the printing telegraph methods employed in cable telegraphy to radio purposes. In general, only high-grade radio circuits are suitable for these applications. In order to make the best use of the wireless circuit, double channelling in time sequence is employed by making the synchronized selecting apparatus of the cable printer also perform synchronized channel switching. In this way two independent channels are established, which under good radio conditions are operated independently. When conditions deteriorate to the point where the number of errors is too great the channels send the same message simultaneously but with a time interval of 2 sec. between the transmissions. This delay is obtained, not by actual retransmission from the tape but by the storage of the signals passed from the tape transmitter in a condenser storage device. These stored signals are subsequently released into the second channel with the required time-delay. At the receiver the first-channel signals are stored in a similar condenser storage device; on arrival of the second-channel signals the two are automatically compared, and if the required correspondence is found the type printer is set up and operates. In its present stage of development the apparatus can be adjusted to give best readability, either through fading or, alternatively, through atmospherics and noise. Experiments are in progress whereby it is hoped to deal with both conditions simultaneously.

A radio circuit has been set up between London and Ascension which forms part of long cable chains extending to Buenos Aires and Cape Town, the connection being maintained without manual transfer at any point. The establishment of this circuit has necessitated the development of a relaying system which converts the 3-element cable code to the normal 2-element (on/off) used in wireless telegraphy. This has been successfully carried out and the circuit is now part of normal world communications.

Reduction of Fading

Frequency wobbling of short-wave transmission to reduce fading has been further developed. This method has advantages over amplitude modulation in that: (*a*) The power of the transmitter is not reduced. (It is necessary to reduce the power of a given telegraph transmitter when amplitude modulation is applied, since the usual limitations governing telephony transmission apply.) (*b*) The spectrum of the radiated wave shows that the energy is more concentrated in the bands adjacent to the carrier, and consequently less sideband interference is produced.

For the purpose in view it has been found that a suitable modulation is provided with minimum sideband spread if the frequency of the shift is at 400 cycles per sec. with a carrier frequency swing of ± 800 cycles per sec. This method of reducing fading is in use on most of the short-wave telegraph transmitters where spaced-aerial (diversity) reception is not employed at the distant terminal.

One of the cheapest ways of effecting an improvement is by the use of the series phase directive receiving aerial, the design of which has been the subject of study. This type of aerial is now largely used for the reception of telegraph signals. In this connection it is interesting to note that an increase in directivity of the aerial system may result in a reduction in effectiveness of a circuit due to an apparent increase of fading. In explanation it is suggested that these waves undergo large angular deviations at times.

Miscellaneous

The approach of the next sunspot maximum has necessitated the provision of additional plant and aerials to meet the anticipated adverse conditions. These preparations even envisage the possibility of increased use of long-wave circuits to maintain communications.

A permanent record of earth-current intensity is now kept. It has been found that these currents increase about 2 hours before a general fade occurs. This discovery is of importance in that traffic diversions may be arranged to forestall the fade and so minimize the attendent dislocation of the services.

A system has been developed whereby two independent telegraph signals may be emitted from a single transmitter and applied to a number of short-wave telegraph transmitters both on point-to-point services at Leafield and on the long-distance ship service at Portishead. Briefly, this system consists of providing two crystal oscillators in each wave-range, which produce two carrier frequencies some 6 or 8 kc./sec. apart. These two frequencies, by means of a third oscillator, are radiated alternately, under "marking" conditions for periods of approximately 0·0025 sec., and two messages can be sent by individually "keying" the two carrier frequencies. The system has helped materially to reduce delays in handling traffic during busy hours without appreciable addition of transmitting equipment.

(7) LONG-DISTANCE TELEPHONE CIRCUITS

Further radio-telephone services, all using short wavelengths, have been opened to Japan, Kenya, Iceland, and Kuala Lumpur (via Bandoeng); and a temporary service has been opened to Portugal since that country has been isolated telephonically from the rest of Europe during the present trouble in Spain.

Short-wave Telephone Reception

As has already been mentioned, there has been some evidence, during the early part of 1938, indicated by adverse transmission conditions, of the approach of high sunspot activity and magnetic variations; these are expected to reach a maximum during 1939–40. Attention has naturally been turned to methods whereby the resulting dislocation of telephone traffic under disturbed conditions of short-wave propagation may be reduced. A new technique of reception has been developed which, it is hoped, will offset to some extent the bad conditions to be expected, and the British Post Office is co-operating with the Bell Telephone Laboratories in America in this work. One of the most impressive features of the new method lies in the aerial system, which consists of 16 rhombic aerials in line, covering about 2 miles in the direction of reception. Each aerial is connected individually to the receiving station by an underground coaxial copper-tube line. By adjustment of the relative phases of the high-frequency currents induced in each antenna it is possible to produce a sharply defined cone of reception, the axis of which can be directed in both the horizontal and vertical planes. By this means the circuit degradation due to interference between rays arriving by different paths, as evidenced by both fading and speech distortion, can be reduced.

One telephone channel to America, working on short wavelengths, has been operated on traffic on the single-sideband reduced-carrier system for some time, and has given satisfactory performance. The expected improvement in signal-to-noise ratio has been generally obtained, but changes in the levels of the output from the receiver under conditions of deep fading were found to be accentuated compared with the normal double-sideband system. To overcome this a " constant volume " amplifier has been added to supplement the automatic gain control of the receiver. The constant-volume amplifier is connected between the receiver output and line and, over a wide range of input levels, supplies a level to line which is constant within ± 2 db.

Miscellaneous

Experiments whereby two separate conversations may be transmitted by using two separate single sidebands, one on either side of the nominal carrier frequency, have been in progress for some time and give promise of being very successful. The possibility of doubling the number of available channels by this method of working has obvious advantages in view of the congestion of the short-wave frequency spectrum, and also allows the introduction of improved secrecy systems.

Improvements to the equipment at the radio-telephony terminal have been made in an endeavour to improve the service to the subscriber. The expander portion of a " compandor " has been modified to act as a " noise reducer " by a slight alteration of its characteristic, whereby incoming signals below a certain arbitrary adjustable level, i.e. chiefly noise, are attenuated, and signals above the arbitrary level—chiefly speech signals—are passed without loss. This apparatus has been used very successfully on some difficult circuits, but it must be used judiciously since it further accentuates fading. On circuits where fading is not too severe an increase in the commercial grade of circuit is generally obtained.

To improve the control of the level passed to the transmitter, which is usually carried out manually, an " automatic technical operator " has been developed and tested successfully. This device performs the function of passing a constant level of signal to the transmitter irrespective of the variations, within wide limits, of the incoming level from the subscriber.

(8) ULTRA-SHORT-WAVE CIRCUITS

Increasing use has been made of frequencies above 50 Mc./sec. to provide communication between points separated by water where the distance does not exceed about 100 miles; and several islands round the coast have now been provided with telegraph and telephone circuits by radio equipments using these frequencies. Transmitters, of power output varying between 1 watt and 250 watts, depending on the range required, have been developed for working on these frequencies with the carrier frequency controlled by quartz, or in some cases tourmaline, crystals. Superheterodyne receivers, having the first beating oscillator frequency controlled by a suitable quartz crystal, have also been developed simultaneously for the services. In cases where more than one circuit is required, apparatus has been developed whereby two speech channels may be simultaneously impressed upon a common radio-frequency carrier, and such a system has been in successful operation for over 2 years to provide telephone circuits between the mainland and the Channel Islands. In parallel with these developments a 9-channel multiplex ultra-short-wave equipment has been installed for working between Scotland and Ireland across the North Channel, and in this equipment all nine speech channels are imposed upon one radio-frequency carrier.

Telegraph circuits using a voice-frequency teleprinter working over ultra-short-wave radio links have been provided to the Scilly Islands, and will shortly be provided to several islands off the coast of Scotland.

(9) MILITARY EQUIPMENT

Progress in military equipment has lain principally in the application of better methods and materials which have become available with the advance of the art. Examples of better methods are the use of voice-switching in telephone apparatus and automatic listening-through devices for telegraphy. Materials such as iron-dust cores and ceramics with special properties have enabled the size and weight of the sets to be reduced and the robustness and reliability to be increased. It is interesting to consider three main types of equipment, namely medium-power pack sets, small-power man-pack sets using batteries, and tank sets.

Medium-power Pack Sets

The power supply for these sets is either a pedal-driven or a light-engine-driven generator. A very simple form of voice-operated carrier suppression for telephony is now incorporated and the same device is used for listening-through on telegraphy. By "listening-through" is meant the ability to hear the distant station in the intervals between the elements of the morse characters being sent by the local transmitter. Inter-tuning is also provided, which enables the receiver to be set to exactly the transmitter frequency and vice-versa. The transmitter circuits are completely ganged, so that any desired wave may be selected by a single adjustment. A separate rod aerial is also employed for the receiver, to permit of continuous watch-keeping when on the move. Pentode valves are used in the transmitter circuits to eliminate neutralizing arrangements with their added weight and complexity. The wave range covered by these sets is generally about 30 to 750 m. The range is usually about 100 miles for telegraphy and 30 miles for telephony.

The receivers are of the superheterodyne type with simplified adjustments. They incorporate all such modern ideas as iron-dust cores, automatic volume control, illuminated scales (which are also provided for the transmitter), drum switching of coils for various ranges, etc.

Small-power Man-pack Sets

Batteries are used as the source of power. The wavelengths are in the region of 100 m. A typical complete pack—transmitter and receiver, aerial, and all supplies—is under 30 lb. in weight. The aerial is an integral part of the set, and the whole equipment can be operated whilst on the march. The transmitter and receiver are tuned simultaneously by one handle, thereby speeding up the process of establishing communication. The sets are arranged to work on either C.W. or telephony, and the aerial power is of the order of 1 watt. These outfits will give a range of about 7 miles on telephony and 15 miles on C.W. telegraphy.

Another still lighter type of equipment for restricted ranges is one operating in the ultra-short-wave band, generally in the region of 5 m. The complete pack weighs about 12 lb. and works on telephony only. Its range is about 2 miles over flat land, but contact has been made up to 20 miles between hills or other elevated positions.

Tank Sets

These sets are arranged to operate from a 12-volt battery which is interchangeable with the tank motor-starter battery. The wavelength employed is about 100 m., and ranges up to 10 to 12 miles are obtained on telephony between aerials permanently fixed on the tanks.

The receivers are provided with visual indicators to facilitate the process of establishing common-wave working on the frequency chosen by the commander tank. Superheterodyne reception is employed and one-knob control is provided, the scales being directly calibrated; very good frequency discrimination is obtained. The complete transmitting and receiving circuits are contained in a box only 15 in. × 11 in. × 7 in., fitted with special shock-absorber mountings.

For small tanks and armoured cars a set somewhat similar to the battery-driven man-pack equipment is employed. These sets are made easily detachable from their mountings so that they may be employed in the field if required.

It is important to note that all military telephone equipments may be used with a special microphone which is strapped to the neck of the operator. This enables him to continue working whilst wearing a gas mask.

(10) RESEARCH AND DEVELOPMENT

Ionospheric Investigations

Considerable additions to our accurate knowledge of the ionosphere have arisen from the development of manual and automatic recording by means of pulses, of the variation of the ionospheric heights with the frequency of the emitted wave. The results yield the heights and maximum ionic densities of the layers concerned, i.e. the E layer at about 100 km., the F_1 layer at 220 km., and the F_2 layer at 250 to 400 km.

Material obtained in various parts of the world has made known the diurnal, seasonal, and year-to-year change in the ionospheric characteristics, as well as latitude changes in some isolated cases. Although the information is by no means complete enough for a world survey, it has been sufficient, with the help of eclipse results, to confirm that the E and F_1 ionization is produced in a fairly regular manner by the sun's ultra-violet light.

The mechanism of the ionization of the F_2 layer is not yet understood. It has been found that the maximum ionization in the layers varies largely throughout the sunspot cycle, being greatest when the sunspot numbers are greatest. This causes the now well-known cyclic change in the optimum waves for long-distance transmission.

The theory of the double refraction of vertically-transmitted wireless waves in the ionosphere has been developed and checked experimentally. It has been found to explain accurately the splitting and polarization observed on downcoming waves. In particular, it has been confirmed that free electrons, and not positive ions, play the predominant part in refracting electric waves in the ionosphere. A great deal of information has also been obtained in the propagation of long waves, both as regards their height of reflection and their downcoming polarization.

Although the theory of oblique (glancing incidence) transmission has not been advanced to such an extent, progress has been made by the formulation of a very important equivalence theorem, according to which the oblique-incidence characteristics can be derived from the vertical-incidence characteristics. The theorem is known to be not quite accurate, but experimental tests made so far agree closely with the theoretical results, so that the inaccuracy cannot be very great. It is now possible to obtain with very fair precision, from the vertical-incidence measurements, the skip frequency for any given distance.

Constitution of the Ionosphere

Recent measurements at vertical incidence of the reflection coefficient of the ionosphere have shown how it is possible to measure the electron collision frequency and hence the specific attenuation, at various points in the layer. It has been confirmed that practically all the loss in long-distance transmissions occurs in passing and re-passing through the E layer, with a resultant attenuation inversely proportional to the square of the frequency. On the purely physical side these collision-frequency results, coupled with the diurnal density variations, have suggested that the upper atmosphere (above 200 km.) is at a very high temperature, estimated to be 1 000° C. The seasonal and diurnal variations of this temperature are not known with any certainty.

High-power impulse technique has revealed the presence of irregular but fleeting ionic clouds in the E region of the ionosphere. They are present both by day and by night, produce scattered signals, and set a limit to the accuracy of direction-finding at distances where the ray reflected from the ionosphere is important. In this connection it may be mentioned that the long and difficult problem of producing a direction-finder entirely free from polarization error has been solved. The spaced-frame system, although not yet in a practical form, has reduced these errors to less than the residual errors caused by lateral deviations of the rays in the ionosphere.

Magnetic Storms and Sudden Fades

This review would not be complete without a reference to the investigation of magnetic storms and sudden fades which cause such disastrous interruptions to various services. The poor conditions in high magnetic latitudes, associated with magnetic storms, have long been known. It has now been found that the effects are associated, in general, with a reduction of ionic density in the F_2 layer, and an increase of attenuation in the E layer. The completely different character of another type of sudden violent fade, starting very abruptly and lasting generally for minutes rather than hours, has now been recognized. These fades occur only when the transmission path is wholly daylight. Convincing evidence has been obtained to show that these fades, which are often accompanied by an increase of noise, are associated with bright hydrogen eruptions from the sun, which presumably produce violent ultra-violent emission which penetrates below the E layer, producing excessive attenuation. These fades have become much more numerous with increasing sunspot activity.

Transmission Theory

The theory of transmission over a flat, imperfectly conducting, earth has been refined, and some of the original difficulties and obscurities have been cleared up. In particular, the theory of ultra-short-wave ground-ray propagation has been extended to keep pace with television developments. The complete theory of the diffraction of waves round a spherical earth of any conductivity has been treated by three methods which are all in substantial agreement.

These theories include the effect of raising the transmitter and receiver above the ground, and one of them allows the effect of refraction to be estimated. Measurements have shown that the signal intensity at places beyond the visual range is greatly in excess of the calculated values, and this implies that there is considerable refraction in the lower atmosphere.

In this connection, measurements with very short impulses have shown that weak reflections corresponding to equivalent heights of 5 to 60 km. sometimes occur. Their origin is as yet unknown. They may be responsible for some of the effects observed in ultra-short-wave transmission beyond the visual range.

Frequency Control

One of the most remarkable developments of recent years has been the universal improvement in frequency stability of continuous-wave transmitters of all types, varying from the very high-power broadcasting station to the small mobile or portable equipments of a few watts output. This improvement has been fostered by the Regulations resulting from the C.C.I.R. Recommendations relating to frequency stabilities and tolerances established at the various international conferences. These Recommendations and Regulations have stimulated work on methods of frequency control, resulting in the accommodation of more and more services in the allotted bands. This closer packing of the channels has again added impetus to this branch of research, on account of the prime necessity for avoiding mutual interference.

The methods of control now employed vary with the type of service, and the transmitters fall broadly into two classes: (i) Those required to transmit one or more fixed spot frequencies. (ii) Those required to cover a wave band and to be able to transmit any selected frequency within the band.

The requirements of the spot-frequency transmitter have proved to be best met by some form of electromechanical primary oscillator. In the early days various forms of valve-maintained tuning forks, steel-rod oscillators, etc., combined with suitable frequency multipliers, were employed. During the last few years, however, the simplicity and directness of the piezoelectric quartz resonator (which can be designed to have a fundamental frequency as high as, for example, 20×10^6 cycles per sec. and to oscillate reliably at this high frequency when maintained by a suitable valve circuit) have won the field for this type of frequency control.

Studies have been made of the high-frequency properties of sections cut at various orientations to the natural axes of the virgin crystal. These investigations have resulted in the establishment of definite rules in regard to the cutting of sections having a substantially zero coefficient of change of frequency with change of temperature. It has been found that various angles of cut and shape of section have special advantages in different regions of the radio-frequency spectrum.

Other investigations have revealed the methods of section whereby crystals having a single major response frequency can be produced. One of the troubles has been the presence of various modes of oscillation in quartz plates. The resonator will oscillate in such alternative modes as will produce the required piezoelectric reactions. It is now possible to choose the section in such a fashion as will accentuate the desired oscillation and so make it certain that the correct frequency of oscillation is always obtained.

All the foregoing work has rendered it necessary to have means of determining with accuracy the principal axes of the natural quartz. By examination of prepared lumps of the original crystal under polarized light, together with the inspection of an etched surface microscopically, the positions of the axes can be determined very accurately. By these means it is now possible to cut slices at predetermined angles in the sure knowledge that they will display the desired characteristics when finished.

Study has also been made of the phenomenon of electric " twinning " in quartz. It has been found that this defect is clearly revealed on a visual inspection of an etched surface at right angles to the optic axis of the crystal, when the two angles of crystallization, together with their random boundary surface, are clearly visible.

By the careful application of this work, combined with a study of mountings in relation to the nodal points of the section when vibrating, it has been found possible to produce a stability of frequency of oscillation of the order of 1 in 10^8. It is interesting to note that this stability may be compared with an accuracy of timekeeping of a clock to within $\frac{1}{3}$ sec. per year.

Of course, this high stability is not required in normal transmitter-frequency control, where a steadiness of a few parts in a million is all that is required. For this purpose the quartz sections are mounted in very simple holders which will plug into suitable sockets, and the cost is remarkably low. The low cost renders their use possible in all spot-wave transmitters of normal type.

For transmitters of other types the adjustable inductance-capacitance oscillatory circuit is still used. Much work has been done, however, in the experimental study of various types of temperature-compensated inductance coils and condensers. The suggestions and experiments which have been made are legion. Suffice it to say that although stability equal to that given by the quartz oscillator under varying temperature conditions has not been achieved, nevertheless the inductance-capacitance oscillator has reached a stage where its performance is satisfactory in many cases where a continuously adjustable frequency is required in some specified band.

The aim of all the foregoing work is principally to remove the necessity for thermostatic control of the primary oscillator. This temperature control has been a major difficulty in the design of frequency-stabilizing apparatus for small and mobile stations. The removal of the necessity for its use is no mean achievement.

Adcock Direction-finder

The increasing use of the Adcock direction-finder has revealed errors which experience with the normal rotating frame direction-finder would not lead one to expect. Study has shown these errors to be due to lack of uniformity in the electrical properties of the soil in different parts of the site of the Adcock aerials. Careful work has shown that a single vertical conductor does not exhibit the same effective height for waves incident in different directions. Also, of course, the four aerials, although accurately and identically symmetrical physically, may have electrically different heights. In general these effects are not large enough to be of major importance; but on very difficult sites they may have a controlling influence on the usefulness of the station. Methods of correction are at present the subject of study.

The design of radiogoniometers for use in all types of fixed-aerial direction-finders has been much improved. This improvement has been principally due to the development of better methods of measurement and measuring apparatus, especially in the very high-frequency regions.

The progress of direction-finding has led to the development of multi-channel direction-finders which operate via long buried cables connecting the aerial system with the radiogoniometers and receivers. These multi-channel equipments will now work with complete absence of interference between channels and with an accuracy of better than ± 1° on all channels.

Signal-measurement Apparatus

Improvements in ultra-high-frequency signal-measuring apparatus are also to be recorded. Apparatus has now been developed working from 15 m. down to 2 m. The apparatus is simple to use and is self-checking. A part of this work has involved the investigation of the behaviour of thermocouples with currents of very high frequency. Errors of as much as 50 % have been observed, and new designs of apparatus for current measurement have resulted.

Work on signal-measurement apparatus in the band from 14 m. to 2 000 m. has been directed towards the simplification and improvement of practical engineering measuring instruments. Direct-reading signal-measuring equipments in this band now exist which will measure signal strengths from $\frac{1}{4}$ microvolt to 1 volt per metre.

Commercial-service Receivers

Steady progress has been made in the improvement of commercial telegraph and telephone receivers. The improvements consist principally in the careful and scientific application of such components as iron-dust cores (toroidal in shape), crystal band-pass filters, and improved automatic frequency-control (self-tuning) apparatus, to receivers. The stability and reliability which this considerable work has secured is such that it is now possible to switch on a complex short-wave telegraph or telephone commercial receiver at any desired moment with the certainty that the circuit will operate at its best without any preliminary tuning or adjustment whatsoever, and to leave it running unattended for as long as the circuit is open. These improvements have greatly simplified and reduced the cost of running a commercial receiving station.

Vacuum Tubes

The development of television has led to much work on tubes for use in pick-up cameras, and, in parallel, large cathode-ray oscillograph tubes with much improved electrode systems have been produced. This aspect of the subject is dealt with in another review.

Electron Multiplication

An interesting application of the phenomenon of secondary electron emission, long familiar in the "dynatron" valve characteristic, has been found in the development of the "electron multiplier." Broadly, the principle lies in the fact that it is possible to liberate five or more electrons from a target surface by the impact of one high-speed electron. By disposing a collector electrode maintained at a suitable positive potential near to the target the secondary electrons may be made to fall on it. In this case the current arriving at the collector anode may be several times greater than the current in the original electron stream impinging on the target. The process may be repeated a number of times, and so a very large magnification of the original current may be obtained. The method is especially applicable to the magnification of the small currents from photo-electric surfaces. The photo-electrons may themselves be accelerated and form the primary electron stream to be multiplied. In this way Johnson noise in the input circuit of the usual photocell plus thermionic-valve amplifier may be avoided. Electron multiplication has not so far proved an advance on more conventional means for the amplification of the currents in wireless receiving circuits. Probably the fundamental reason for this lies in the fact that most of the noise in a receiving circuit originates in thermal agitation of the electrons in the input circuit.

"Beam" Valves

Another important development has been the "beam" valve. By using an electrode system somewhat similar to that of a cathode-ray oscillograph it is possible to produce a flat ribbon-like beam of electrons which can traverse a space and then fall on a tiny target maintained at a suitable positive potential. The current intensity in the beam can be controlled by the usual control electrode. By adopting this construction, valves having very small electrode capacitances can be made, and, what is more important, the inter-electrode capacitance between anode and control grid is extremely small, being of the order of only 1 per cent of that of the normal screen-grid valve. The valve lends itself to good screening design, and thus very perfect isolation between input and output circuits can be secured.

The use of a similar principle is shown in the design of the "aligned grid" electrode assembly now used in most multi-electrode valves. In this construction the control, screening, and other grids are so made that the constituent wires of successive grids lie directly behind each other in the path of the electrons from cathode to anode. Thus, the wasted currents to the various grids are reduced since successive grid wires lie in the shadow cast by those nearer the cathode. This method of construction is of special importance in valves for transmitters and output power valves for speech amplifiers.

It is impossible in the course of this brief review to deal with the many-sided activities of research in this field; the foregoing examples are selected from the large amount of work going on in vacuum-physics laboratories both at our universities and in commercial undertakings.

High-frequency Cables

The development of multi-channel carrier telephony and the engineering problems arising from the necessity, in some cases, of transmitting and receiving from aerial systems remote from the station buildings, have led to extensive studies followed by the development of various types of high-frequency cable. The considerations are: (1) that the loss in transmission shall not be excessive, and (2) that the constants of the cable shall not be such as to render the design of termination transformers impracticable.

The ideal cable is a copper tube outer with a smaller concentric inner tube supported at intervals by insulators of high resistance and low dielectric loss. The working dielectric is air or other suitable dry gas. Cables, or "feeders," of this type are technically the best but are expensive to make, install, and maintain.

Insulating Materials

As a result of the study of insulating materials it is now possible to construct semi-flexible high-frequency cables. It has been found that certain thermo-plastic polymerized compounds having a styrene base have the necessary dielectric properties, and are flexible. Many designs of cable have been evolved, but in general the insulation is formed as a thread which is wound round the core, and the whole then has a lead sheath extruded over it to form the outer concentric conductor.

Improvements in ceramic insulating materials have also rendered it possible to adopt designs utilizing freely this type of insulation in the form of beads or discs. Some of the new ceramics have a dielectric loss not much greater than that of air.

(11) ACKNOWLEDGMENTS

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BROADCASTING AND TELEVISION*

By Sir NOEL ASHBRIDGE, B.Sc.(Eng.), Member.†

PART 1. SOUND BROADCASTING

General Development

During the period under review—from the end of 1934 to the end of 1938—the number of licence-holders in European countries for which reliable statistics are available has increased by more than 60 %. During the first 3 years of this period there was an increase of 50 %, the figure for 31st December, 1937, being 31 738 120. Under the Lucerne Plan of 1934 there were 254 broadcasting stations in Europe, the total power being 4 300 kW. The mean power of these stations was about 17 kW and the highest power radiated by any transmitter was 500 kW. At present there are about 280 medium- and long-wave stations in Europe, with a total power of more than 8 100 kW. The maximum power of any station is still 500 kW, but the mean power has increased to about 29 kW.

In this country the number of licence-holders has increased from 6 747 340 at the end of December, 1934, to 8 908 366 at the end of December, 1938. The scheme of broadcast distribution has steadily progressed and it is now estimated that 98 % of the population of Great Britain and Northern Ireland can receive one programme satisfactorily both by day and by night; furthermore, 84 % of the population have a choice of two or more B.B.C. programmes. This result has been achieved by the construction of a number of new transmitters. The long-wave station at Droitwich had already been completed before the end of 1934. It was followed in February, 1935, by a 70-kW medium-wave transmitter on the same site, which covers the Midland Region and replaces the old Regional transmitter at Daventry. In March, 1936, a 100-kW medium-wave transmitter was put into operation at Lisnagarvey to provide a Regional programme service for Northern Ireland and to replace the low-power transmitter at Belfast. In October of the same year a 60-kW transmitter was opened at Burghead to serve the North of Scotland, and in February, 1937, a medium-power station at Penmon began to provide a Welsh programme service to the Northern part of Wales. In July of that year the Welsh Regional service was divorced from the West of England programme, the Welsh programmes being provided by one of the two transmitters at Washford (and by Penmon in North Wales) and those for the West of England by the other transmitter at Washford and by low-power transmitters at Plymouth and Bournemouth. A high-power transmitter at Stagshaw was opened in October, 1937, to serve the North-East of England, replacing the low-power transmitter at Newcastle. A new medium-power transmitter at Redmoss near Aberdeen replaced the old Aberdeen transmitter in September, 1938. In addition to these medium-wave transmitters, several new short-wave transmitters and an extensive system of aerial arrays were constructed on the Daventry site and were first used for the Coronation broadcasts in May, 1937. Programmes in Arabic have been broadcast daily from Daventry since January, 1938, and programmes in Spanish and Portuguese since March, 1938. Work is in progress on a new 100-kW transmitter at Start Point in Devon and a medium-power station at Clevedon, near Bristol, which together will give a Regional programme service to listeners along the south coast from Cornwall to Sussex and an improved service in the south-west of England.

In order to provide increased and improved studio accommodation a new group of studios had already been opened at Maida Vale in October, 1934. Since then new studio centres have been provided at Bangor, North Wales (December, 1935), Swansea (October, 1937), and Glasgow (November, 1938). The Glasgow studio centre incorporates the latest improvements in construction, treatment, and equipment; there are 10 studios, one being the second largest in the country. New studio buildings are under construction in Belfast and in Aberdeen.

Synchronization of Transmitters

Before the beginning of the period under review the increased number of transmitters to be accommodated in the limited number of wavelength channels available in the medium-wave band allocated to broadcasting had already made it imperative to develop methods of synchronization to enable two or more transmitters to work on the same wavelength. It was found that even if two synchronized transmitters always radiated the same programme, the area lying midway between them was poorly served because of interference between the waves received from the two sources. This interference was caused partly by phase differences and partly by small differences between the actual frequencies of the two transmitters. A considerable reduction in the size of this area was effected by increasing the accuracy of synchronization. At first synchronization was effected by transmitting a low-frequency tone over Post Office lines from one station to another, the frequency of this tone being a definite sub-multiple of the carrier frequency of the transmitters. It was multiplied again at the driven transmitters to provide the actual carrier frequency of the group. This method suffered from certain disadvantages: it was necessary to use a network of lines interconnecting the synchronized transmitters, solely in order to carry the synchronizing tone: any slight alteration in the attenuation of these lines caused a severe disturbance in the programme received at a point in the area where the ratio of the field strengths of the two transmitters was not greater than about 5 to 1. The first disadvantage was overcome in some cases by superimposing the driving tone on the line which carried the programme, but the frequency of the driving tone had to be chosen so as to be just within the band transmitted by these lines; at the same time filters had to be provided at the receiving end to exclude the tone from the programme. Special precautions, such as the use of tuning-fork filters and oscillators, were taken to minimize the effect of line variations and induced line noise.

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* A review of progress. Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 380.
† British Broadcasting Corporation.

There has, however, been a steady improvement in the frequency stability of individual transmitters. This has been brought about by the growing use of temperature-controlled quartz crystal oscillators. The following Table, based on measurements made at the Brussels checking centre of the U.I.R.,* shows how great the improvement has been:—

FREQUENCY STABILITY OF EUROPEAN BROADCASTING TRANSMITTERS

Maximum variation from mean frequency, c./sec.	Number of stations in:				
	Aug. 1934	Aug. 1935	Aug. 1936	Aug. 1937	Aug. 1938
0 – 1	3	7	20	21	25
1·1–2	3	10	17	16	27
2·1–5	9	23	26	46	26
5·1–10	8	17	16	21	32
0–10	23	57	79	104	110
0–50	42	93	114	138	143

This improvement in frequency stability has made it possible to dispense with the driving tone formerly used for synchronizing and to drive the individual transmitters of a synchronized group by independent quartz crystals. It is, however, still necessary to transmit a synchronizing tone from one transmitter to another for short periods each day so that the frequency can be checked and, if necessary, restored to the correct value. The Welsh Regional and Penmon transmitters have used this method with success for more than a year, and more recently the London, North, and Scottish National group has been changed from tuning fork to independent quartz-crystal drives. The stability of the master crystal at the Welsh Regional station has been found to keep within 0·5 c./sec. over a period of some months—on a carrier frequency of 804 kc./sec. With modern methods of synchronization the ratio of wanted to unwanted signal can be as low as 2 to 1 in field strength.

Design of Transmitting Stations

Before the beginning of the period under review the general increase in the power of transmitters had made the cost of electrical energy a factor of economic importance. Efforts have, therefore, been made to increase the ratio of the output power to the total power consumed by the station. A simple way of achieving this object is to vary the amplitude of the carrier in accordance with the depth of modulation; this system has the effect of operating the automatic gain control on receivers.

Many proposals for high-efficiency modulation systems have been made, and the following are amongst those which have received the most attention—the "High Power Class-B," the "Doherty," the "Chireix," and more recently the "Fortescue," systems.

In the first of these the modulator consists of a high-power low-frequency amplifier with two valves or groups of valves working in push-pull and biased so as to work in the Class-B condition. The average H.T. current thus varies with the depth of modulation, so that the average power consumption is considerably less than the peak power consumed. The modulated valves are driven in such a way as to give maximum efficiency, and modulation is applied to the anodes of the modulated valves through a transformer.

A disadvantage of this form of Class-B system is that a very large l.f. transformer, a large choke, and a large condenser, are needed to carry the high power in the output of the modulator stage. In the Fortescue system this necessity is eliminated by using two banks of h.f. valves in the final stage, one carrying unmodulated high frequency and the other giving an h.f. output proportional to the programme input and working on the Class-B principle. By means of a special impedance-inverting network, the outputs of these two banks are combined to give a normal modulated output. In effect, one bank provides the carrier and the other the sidebands. The bank which supplies the carrier operates at maximum efficiency with a h.f. input, while the other which supplies the sidebands, being a Class-B stage, also works economically.

In the Doherty system there are also two banks of valves, both of which are driven by a normal modulated valve. The first, however, is driven in such a way that in the modulated state the anode peak voltage is almost equal to the H.T. voltage, so that the valves operate at high efficiency. On one half of the modulation cycle the valves amplify linearly, but on the other they limit severely. The other bank is so biased that it is inoperative until the first is beginning to limit, after which it supplies the current necessary to produce a normal modulated output. An impedance-inverting network is again used in order to combine the outputs of the two banks. The system relies upon the balance of the curvature due to the limitation of the valves in the first bank against that due to the bottom bend of those in the second bank.

The Chireix system uses two banks of valves in push-pull, to the grids of which (in series) the modulated h.f. voltage is applied. In addition, a slightly modulated h.f. voltage is applied to the two grids in parallel. This "pilot" voltage does not appear in the output, but by suitably arranging the relative phases it is possible to ensure that the output voltage of each bank is nearly constant throughout the modulation cycle and of such a value that the valves work at maximum efficiency. The output from the two banks is combined in such a way that the two phase-modulated outputs are made to provide a single amplitude-modulated output.

The Class-B system of high-efficiency modulation is in use at the new transmitters at Stagshaw and Redmoss and in two of the new short-wave transmitters at Daventry. In the case of Stagshaw the ratio of the h.f. power output to the total power consumed by the transmitter is about 0·325, compared with about 0·22 for the older system of modulation at low power levels.

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* The International Broadcasting Union—L'Union Internationale de Radiodiffusion.

The Doherty system of modulation is in use for two 50-kW stations of the Canadian Broadcasting Corporation, while the Chireix system is used in several broadcast transmitters on the Continent, notably in France.

The question of aerial design for medium-wave stations has also become more important, since the range of broadcast transmitters working on high power is usually limited at night by the interference produced between the directly propagated wave and the indirect wave reflected from the ionosphere, rather than by the weakness of the received signal in comparison with the level of interference at the receiver. It is, therefore, necessary to endeavour to increase the effectiveness of the propagation of the direct wave and to reduce that of the indirect ray. Mast radiators having a height equal to about 0·55 times the wavelength had proved efficient in this respect before the beginning of the period under review. Masts of slightly less height can be made effective by the addition of a metal ring at the top, the capacitance of which tends to make more uniform the current distribution up the height of the mast at the frequency of the carrier. Mast radiators have been constructed in this country at Lisnagarvey, Burghead, Penmon, Stagshaw, Aberdeen, Start Point, and Clevedon, and they are in use at most of the recently constructed stations abroad. At Budapest the highest mast of this type in the world has been in use for some years. Its maximum height is 1 030 ft. and it is intended for a carrier wave of 546 kc./sec.

Improvements have also been made in the design of the feeder lines which connect the transmitters to the aerials and in the coupling circuits at each end of the feeder lines. The present practice of the B.B.C. is to use concentric copper-tube feeder (supported above the ground) for medium- and long-wave stations, and open-wire feeders for short waves.

Further improvements in transmitters have included better protection against lightning and the use, for short waves, of variable-inductance tuning.

Studio Design

During the period under review certain changes have been made in the method of acoustically treating the internal surface of studio walls. It is still the general practice in this country to make the structure of the walls as solid and non-resonant as possible by using thick brickwork and avoiding steel reinforcement, which would tend to transmit sounds from one room to another. But, whereas it was frequently the custom to cover the wall surfaces with " building board " firmly fixed to them, a number of other materials are now more favoured as a means of obtaining a satisfactory reverberation time as a function of frequency. Improved methods of measurement have shown that building board is not absorbent enough at the lower end of the audible scale, although this effect is sometimes masked by unavoidable resonances in the structure. Panels of wood or lath and plaster have been used for several years, but have recently become more generally popular. These panels are fixed to wood battens on the solid walls in the normal manner, and their resonant properties cause them to absorb sound energy at the lower frequencies. This treatment is frequently combined with ordinary hard plaster, rock wool, and building board, the proportions of the surface covered by the different materials being adjusted to give the desired reverberation time characteristics.

It is now generally agreed that the distribution of these materials in the studio is important, and it is usual in orchestral studios to concentrate the less absorbent surfaces at the end of the studio in which the orchestra is placed and the more absorbent at the opposite end. Practical experience in the use of serrated surfaces for the walls and ceilings of studios has shown that such shapes do in fact tend to eliminate standing waves and to ensure a uniform sound field. The cost of elaborate shapes is, however, excessive and the result of doubtful benefit. Some breaking-up of wall and ceiling surfaces appears to be desirable, but studios with flat, non-parallel walls have given results almost identical with those obtained from rectangular ones. It has, however, been found desirable to provide a tiered platform at one end of each orchestral studio; this not only enables the microphone to pick up the sound waves from each instrument directly without interference from obstructions, but can be made to provide useful acoustic absorption at the lower frequencies with reflection at the higher frequencies. Incidentally it provides a better arrangement of the orchestra from the conductor's point of view. Floors of hardwood strips on battens are also used for similar reasons.

Low-Frequency Equipment

Important developments have taken place in the design and use of microphones of the " ribbon " or " velocity " type; so much so that they have become standardized for use in all B.B.C. sound broadcasting studios. Two new types of ribbon microphones have recently been developed. One is a smaller form of the original model for use in outside broadcasts and in studios, and the other is a " lip " microphone for the use of commentators in noisy surroundings.

In Germany condenser microphones are still widely used, and a single-sided form has been developed which has proved especially useful. In the United States the ribbon microphone has been adopted as standard by the principal broadcasting organizations, although moving-coil and piezo-electric microphones are also used.

A new type of audio-frequency volume meter has been designed which indicates the magnitude of rapid peaks of modulation and conforms to an international specification which has been adopted by the U.I.R. In the U.S.A., broadcasting organizations continue to use volume meters which indicate " mean syllabic " power rather than peak values. In this they follow the practice of the telephone companies, which find this type of meter economical and satisfactory for use on long lines both for ordinary telephone and for broadcast transmission. Several types of automatic volume-limiter and volume-compressor have also been developed, both in this country and in the U.S.A. Experience in the use of limiters in practice shows that, by eliminating the danger of over-modulation from momentary excessive peaks, they permit an increase in the mean level of modulation without appreciable risk of audible distortion.

Improvements in low-frequency equipment generally have resulted from the increased use of the principle of negative feedback and from the use of multi-electrode valves. Improvements are being made in the technical arrangements for controlling, supervising, and switching programmes. The increasing popularity and importance of outside broadcasts have necessitated the design of improved portable equipment for amplifying and monitoring programmes at outside broadcast points. The art of handling such broadcasts reached a high degree of complexity and reliability during the series of broadcasts which were undertaken at the time of the Coronation. On that occasion two special control rooms were established, one in Westminster Abbey for the Home and Empire programmes, and another in the Middlesex Guildhall through which the special commentaries to foreign countries passed.

Sound Recording for Re-Broadcasting

Advances made in magnetic sound recording on steel tape were described in a recent paper* before the Wireless Section of The Institution. This system has proved of great value in short-wave broadcasting systems, where it is often necessary to repeat a programme several times within 24 hours for reception at convenient listening periods in different parts of the world. The recording can afterwards be effaced from the steel tape, which can thus be used many times for different programmes.

Improvements have also been made in the technique of recording on metal discs coated with a preparation of cellulose acetate. There are now many variations of this method in use in various countries. A new system of electro-mechanical recording on specially prepared film has been installed in London by the B.B.C. and in the stations of several broadcasting organizations overseas. Improvements in the equipment used for reproducing both acetate discs and ordinary gramophone records have been made by the B.B.C. so as to facilitate the editing of a number of excerpts from different recordings to form a continuous programme. These improvements include the use of parallel tracking pick-up arms, of electro-mechanical devices for automatically lowering the pick-up on to the record at the moment when the control potentiometer is faded up, and of special methods of synchronizing the end of one record with the beginning of another so as to facilitate continuous reproduction.

Exchange of Broadcast Programmes

A number of improvements have been made in the transmission of broadcast programmes over lines belonging to the telephone administrations in this and other countries; but this subject does not come within the scope of this review. The number and quality of programmes exchanged between the broadcasting organizations of different countries over lines or radiotelephone services has very greatly increased during the period, those in which this country has taken part having trebled in number.

Short-Wave Broadcasting

During the period under review there have been outstanding developments in short-wave broadcasting in all the principal countries of the world.

In this country a service was established at Daventry in 1932 after initial experiments with a transmitter installed at the works of Marconi's Wireless Telegraph Co., Ltd., at Chelmsford. At first there were only two transmitters at Daventry, each of 10–15 kW. They were provided with 12 directional aerial arrays and 6 omni-directional aerials. The service at that time was directed to 5 zones, but it soon became apparent that reception of particular transmission was by no means confined to the zone for which it was primarily intended.

In 1933 the service was reorganized on a time basis, there being 5 transmissions each at a time suitable for reception at a convenient hour in a given part of the world. Later the vertical aerials were replaced by high horizontal dipoles. In 1935 the experimental transmitter was transferred from Chelmsford to Daventry to provide a third transmitter there, and a sixth transmission was added, so that the service now operates for about 18 hours out of the 24.

In 1935 it was decided to use aerials having a narrower angle of transmission in the horizontal plane; it was also decided as far as possible to provide simultaneous transmission on at least two wavelengths for each zone. In order to fulfil these requirements, two high-power transmitters were completed in 1937 and one in March, 1938. Two more are under construction, intended primarily for a service in foreign languages. Thus 6 transmitters are at present working at Daventry, 3 of 50 kW, 1 of 20 kW, and 2 of 10 kW; the two new ones which are under construction will be ready for use early in 1939 and then 5 transmitters at Daventry will each be capable of an output of 100 kW, although they will actually work on 50 kW in accordance with the licence granted by the Postmaster-General.

The design and construction of a new aerial system consisting of 25 different arrays was completed in 1937, and an open-air switching frame was provided to give increased flexibility in the interconnection of the aerials and transmitters. A central drive room is at present being installed to accommodate all the primary frequency standards required for the station and their associated equipment. The output from any one crystal unit can be connected to any transmitter at will.

The service was fully described in a paper by Messrs. L. W. Hayes and B. N. MacLarty which was read before The Institution during the present session.

The study of propagation conditions, which is of fundamental importance to a successful short-wave service, has been assisted by the inauguration of a reporting scheme by which selected observers in each of the regions to be served sends a weekly cable describing reception conditions. These cables, together with the comprehensive log sheets sent by a larger number of listeners, enable seasonal changes in wavelength to be made at the most opportune moment according to the conditions prevailing in the ionosphere.

General Growth of Short-Wave Services

Similar developments have been going on in other countries, notably Germany and Italy, in both of which countries the radiating systems are in general similar to those at Daventry and consist of stacked horizontal dipoles.

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* A. E. BARRETT and C. J. F. TWEED: *Journal I.E.E.*, 1938, vol. 82, p. 265, and *Proceedings of the Wireless Section*, 1938, vol. 13, p. 73.

At Zeesen, the German short-wave broadcasting station, there are 8 transmitters each giving a power output of 50 kW, whereas at Prato Smeraldo, the Italian station near Rome, there are two 100-kW transmitters of similar design to those at Daventry, one 50-kW transmitter similar to those at Zeesen, and two other 50-kW units.

At Zeesen, the feeder system between the transmitters and the aerials consists of articulated concentric cable, while the aerial switching system consists of a motor-operated commutator which gives rapid means of selecting the required aerial array and connecting it to the required transmitter. There have been developments in many other countries, notably in France, the United States, and Holland, while in most European countries some form of overseas short-wave broadcasting is now in regular operation.

At Huizen in Holland a novel form of aerial is in use in which the whole system of two self-supporting masts and the stacked vertical dipole aerial array carried by them is mounted on a turntable on rails so that the beam can be oriented in any desired direction.

Frequency Measurement

In order to ensure the success of international plans for the allocation of broadcast wavelengths there must be an accurate check on the frequency of each transmitter. There are official centres for the measurement of the frequencies of wireless transmitters generally, such as that maintained by the General Post Office at Baldock and that operated by the Bureau of Standards in America. In order to supplement these services in the field of broadcasting and to encourage its members to strive for greater accuracy in the frequencies of their stations, the International Broadcasting Union has maintained a measuring centre at Brussels for a number of years. In 1938 this was moved to a new building specially designed for the purpose and comparatively free from electrical interference. This centre, by virtue of the recognition of the U.I.R. as an " expert " body at the Lucerne and Cairo Conference, works in close co-operation with the administrations concerned in all countries. The precision of measurement at the U.I.R. station is now 2–3 parts in 10^7, and about 600 measurements of long-, medium-, and short-wave broadcasting stations are made every day. In case of any serious departure of a station from its normal frequency the director immediately warns the broadcasting organization concerned.

This U.I.R. checking station is a practical example of useful international co-operation, for the money for its building and equipment has been subscribed by the broadcasting organizations of about 25 countries in the European region and by one of the American networks.

A similar task is performed by the B.B.C. receiving station at Tatsfield which keeps a particularly close watch on B.B.C. stations and others which may interfere with them. Two sets of accurate measuring apparatus are in use. Each of the two basic standards consists of a temperature-controlled quartz-crystal oscillator; the newer one, which has recently replaced an older model, gives a single frequency of 1 000 kc./sec. and has a computed accuracy of ± 4 parts in 10^9. This frequency controls multi-vibrators which produce sub-multiples 1 000 c./sec. apart. A low-frequency oscillator is used, when necessary, to interpolate between these intervals by producing beats with the audible note produced by the measured carrier and the nearest sub-multiple of the standard. The accuracy of the frequency standard is compared every 24 hours with Greenwich time-signals by means of a synchronous clock driven by one of the component frequencies derived from the multi-vibrators. Measurements are made on all B.B.C. long- and medium-wave transmitters three times a day, on all B.B.C. short-wave transmitters at least once during every transmission, and on all European medium- and long-wave transmitters once a week. In addition, between 100 and 200 measurements of short-wave stations overseas are made every day, and special checks are made on complaints of interference. Close co-operation is maintained with the Post Office and with the checking centre of the U.I.R. in the elimination of interference between broadcasting stations. The maximum error of measurement at Tatsfield has decreased during the period under review from about ± 1 part in 10^6 to about ± 2 parts in 10^7. In favourable cases the accuracy may be that of the standard, i.e. ± 4 parts in 10^9.

Reception

At the beginning of the period under review the use of the supersonic-heterodyne principle was restricted to the more ambitious receivers, whereas to-day some of the least expensive models incorporate it. Similarly, provision for reception in the short-wave band was not widespread in 1934, and most listeners desiring to receive short-wave programmes used convertors or adaptors. To-day, however, most receivers include at least one short-wave band, many covering reception of the 14-metre (21·5 Mc./sec.) band or even of the sound channel of the London television station (41·5 Mc./sec.). The increasing interest in short-wave listening is indicated by the growing use of receivers of the so-called " Communications " type.

Automatic gain control has become practically universal, whereas automatic tuning correction is still only comparatively rarely included.

A development which has shown a remarkable rise in popularity is " push-button tuning "; in fact, it is interesting to observe that its popularity, as represented by the number of models upon the market, has increased much more quickly here than in the United States. The three systems most commonly used are:—

(*a*) Electric motor drive of the usual multi-unit tuning condenser.
(*b*) Selection by means of pre-set tuning condensers.
(*c*) Selection by means of pre-set tuning inductors.

Ease of tuning has received considerable attention, quite apart from " push-button " operation, and several systems of " band-spreading " are now current practice. Tuning by varying the permeability of the cores of inductances, instead of by variable condenser, is now frequently used.

More attention is now paid to quality of reproduction than formerly, although it appears to be difficult to improve the results obtained from the best examples of table receivers, which are admittedly less satisfactory than those having larger cabinets. Two or three years ago there was a tendency to use two or more loudspeaker units to provide an adequate output throughout the desired range of frequencies, but this requirement can now be met by the use of single loudspeakers of improved design.

The efficiency of output valves has been improved and many receivers are available with single output valves for which an undistorted output of 5 watts is claimed. Signal-frequency amplifying stages are now commonly used with supersonic heterodyne receivers, and for this purpose the special types of tetrodes and pentodes developed for use in television receivers are particularly suitable.

Electrical Interference

A paper* read before The Institution by Messrs. Gill and Whitehead set forth the present state of the campaign against electrical interference. Briefly, the work which has so far been carried out has enabled the British Standards Institution to formulate specifications for the tolerable limits of interference on medium waves and long waves, for the components of interference-suppressing devices, and for the design of apparatus for measuring the strength of interfering voltages and interfering fields. Work is continuing on the suppression of the interference caused to short-wave broadcasting and to television by electro-medical equipment and by the ignition systems of motor-cars. Methods have already been elaborated for reducing the interference so caused. Considerable attention has also been paid to the design of receiving equipment so as to reduce its susceptibility to interference.

International Regulations Affecting Broadcasting

The International Telecommunications Conference which met in Cairo in February and March, 1938, has approved a number of changes to the current radiocommunication regulations drawn up at the Madrid Conference in 1932. The allocation of wavebands to the various kinds of radio services has been revised, and the new allocation should take effect from the 1st September, 1939.

No change has been made in the so-called " long wave " broadcasting band, which remains as 160–265 kc./sec. (1 875–1 132 m.), with the existing restriction that the location of stations using frequencies at the higher-frequency end (240–265 kc./sec.) should be in the more outlying parts of Europe. The medium waveband has been extended at the higher-frequency end by 60 kc./sec., becoming 550–1 560 kc./sec. (545–192·3 m.), giving some 6 additional broadcasting channels. Receivers covering the new waveband below 200 m. will be necessary in countries to whose stations these channels may be allotted.

Provision has been made in the Cairo Regulations for the first time for broadcasting in the intermediate waveband 150–60 m. Its use is limited to tropical or semitropical regions (between 30° N and 30° S lat.) where the high noise-level due to atmospherics prevents to a great extent the use of the normal medium- and long-wave bands, and where there has been as a result a tendency for stations intended mainly for local service to use frequencies in the short-wave bands below 50 m., intended for long-distance communications only. The new wavebands are:—

2 300–2 500 kc./sec. (130·4–120 m.)
3 300–3 500 kc./sec. (90·9–85·7 m.)
4 770–4 965 kc./sec. (62·9–61·2 m.)

Certain territorial restrictions (especially in America) and limitations of power are, however, imposed.

An additional 500 kc./sec. has been provided for short-wave broadcasting between 50 and 13 m., there having been no change since the original allocation by the Washington Convention of 1927. The wavebands become:—

6 000–6 200 kc./sec. (50–48·39 m.), an increase of 50 kc./sec.
7 200–7 300 kc./sec. (41·67–41·10 m.), a new band of 100 kc./sec. open to broadcasting, except in America, but shared with amateurs.
9 500–9 700 kc./sec. (31·58–30·93 m.) an increase of 100 kc./sec.
11 700–11 900 kc./sec. (25·64–25·21 m.), unchanged.
15 100–15 350 kc./sec. (19·87–19·54 m.), unchanged.
17 750–17 850 kc./sec. (16·90–16·81 m.), an increase of 50 kc./sec.
21 450–21 750 kc./sec. (13·99–13·79 m.), an increase of 200 kc./sec.

The band of 1 000 kc./sec. at about 11·5 m. remains unchanged.

Until the Cairo Conference the allocation of wavebands to the principal services stopped at the higher-frequency end at 28 000 kc./sec. (10·71 m.). With the advent of television and the increased use of ultra-short waves for air navigation and other purposes, the Cairo Conference decided that some regional allocation allowing development to some extent on the same lines in each country was required. An allocation of the ultra-high-frequency bands, to apply throughout Europe, has accordingly been made as far up as 200 Mc./sec. (down to 1·5 m.). The bands allotted for television services are:—

40·5–58·5 Mc./sec. (7·41–5·13 m.), 64–70·5 Mc./sec. (4·69–4·25 m.), 85–94 Mc./sec. (3·53–3·19 m.), 170–200 Mc./sec. (1·76–1·5 m.).

The Madrid Convention of 1932 was followed by a European Governmental Conference at Lucerne in 1933 for the purpose of establishing a wave plan among European broadcasting stations. The Lucerne Plan remains in force, although its allocations in the long-wave band were not accepted by a number of countries, mainly because the number of long-wave stations existing or projected entailed an insufficient separation between them in the waveband available. The existing allocations in the long-wave band mainly follow the temporary plan drawn up by the International Broadcasting Union (U.I.R.) at the end of 1933, which, although in some respects unsatisfactory, has afforded a *modus vivendi* for the existing long-wave stations.

The Cairo Conference issued instructions for the holding of a European Conference to revise the Lucerne Convention, taking into account the changes in the wavebands effected by the new regulations. This is to be concluded in time to allow the plans to be put into force at about the same time as other changes arising from the Cairo Conference. The International Broadcasting Union was charged with the task of preparing, at a meeting to be held at Brussels in the autumn of 1938, a draft plan of allocations for consideration by the European Governmental Conference to be held in Switzerland before the 1st February, 1939.

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* *Journal I.E.E.*, 1938, vol. 83, p. 345, and *Proceedings of the Wireless Section*, 1938, vol. 13, p. 209.

Amongst other provisions the Cairo Regulations have fixed new and more strict standards for the permissible limits of frequency tolerance to be observed in the case of all types of radio stations. The tolerance for short-wave broadcasting stations installed after the 1st January, 1940, has been reduced from 0·01 % to 0·005 % (5 parts in 100 000).

Previous to the Cairo Conference, the fourth meeting of the International Consultative Technical Committee for Radio Communications (C.C.I.R.) was held at Bucharest in July, 1937, and recommendations, some of them revising the " avis " of previous meetings, were made on a number of technical questions of international importance. Those which were of interest from the broadcasting aspect included such subjects as single-sideband working, the use of anti-fading aerial systems, the limitations of harmonics, the propagation characteristics of the various orders of wavelengths, taking account of both seasonal and diurnal variations, and precise figures for the frequency tolerance recommended for application by the various types of transmitters. This last recommendation after some modification was embodied, as mentioned above, in the Radio Regulations approved at Cairo.

PART 2. TELEVISION BROADCASTING

General

The report of the Television Committee appointed by the Postmaster-General was issued in January, 1935, and recommended that the B.B.C. should be responsible for establishing a public high-definition television service using ultra-short waves. Experimental transmissions were radiated in August, 1936, and the Alexandra Palace station was formally opened by the Postmaster-General in November, 1936. A regular programme service has been in continuous operation since that date. At first two different standards of transmission were used, but in February, 1937, a single standard was adopted on the advice of the Television Advisory Committee.

According to this standard (i.e. of the Marconi-E.M.I. television system) the picture is scanned with 405 lines at 50 frames per second, the frames, each of 202½ lines, being interlaced to give 25 complete pictures per second. The peak value of the carrier wave, at 100 % modulation, corresponds to a full-white picture area, while 30 % peak carrier value corresponds to full black. The remaining portion of the carrier, that is from 30 % modulation to zero, is used for the synchronizing signals. This form of modulation is sometimes referred to as " positive " and is in the opposite sense to that used in the U.S.A., where the black level corresponds to about 70 % peak modulation, the white level to zero, and the synchronizing signals to 100 % modulation. The Television Advisory Committee has decided that the London television standards shall remain in force for at least 3 years from the beginning of 1938.

The nominal hours of transmission are now from 3 to 4 p.m. and from 9 to 10 p.m.; there is also a morning transmission intended for the purpose of demonstrations and tests by the radio trade.

The vision transmitter and the studio and control room equipment at the Alexandra Palace station are in accordance with the system developed by the Marconi-E.M.I. Television Co. A number of improvements have been introduced during the two years since the opening of the service; in particular an improved type of Emitron tube, known as the long-gun type, is now in use for studio work, while for the transmission of film the continuous-motion film projector has replaced the type employing the standard shutter type of film projector, with improved results. In the case of broadcasts from places remote from the studio, and in particular for sporting events, a further type of Emitron tube, known as the super-Emitron, has been introduced which makes it possible to obtain satisfactory results with telephoto lenses.

The range of the transmitter is considerably greater than was at first expected, and satisfactory reception is possible up to a radius of about 30 miles except in localities where interference is very severe. In exceptional cases reception has been reported at distances up to 200 miles.

Transmitters

The system of transmission was described in papers* read before The Institution last session. The vision transmitter gives a peak output power of 17 kW, and the sound transmitter is rated at 3 kW carrier power or 12 kW peak power at 100 % modulation. The vision transmitter has linear frequency and phase characteristics up to 3 Mc./sec. The carrier frequency of the vision transmitter is 45 Mc./sec, and that of the sound transmitter 41·5 Mc./sec.

Studios

A second studio at Alexandra Palace has now been equipped as a separate unit with its own control room and producer's control gallery. The construction of a third and much larger studio is under consideration. This studio will have 5 main stages and a control tower, from which the producer and the engineers in charge of microphone-mixing and camera-mixing will have a clear view of all the stages.

A new central control room has been equipped with fading and monitoring apparatus to assist in maintaining continuity between the various parts of the programme which come from different sources. To this room will come programmes from the three studios and outside broadcasts which arrive at Alexandra Palace either by way of the balanced-pair television cable or by the ultra-short-wave link.

The central control room has also been equipped with timing-pulse generating equipment, which is used to synchronize the control pulse generators associated with each of the three studios. An advantage of this arrangement is that it becomes possible to mix the vision output of the three studios.

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* *Journal I.E.E.*, 1938, vol. 83, pp. 729 and 758, and *Proceedings of the Wireless Section*, 1939, vol. 14, pp. 36 and 65.

Outside Broadcasts

A great deal of attention has been devoted to television outside broadcast equipment. Three different methods are available for conveying the vision signals from the outside broadcast point to the transmitter at Alexandra Palace, and a combination of more than one of these methods is sometimes necessary. Two complete mobile units are available to provide an ultra-short-wave link working on a frequency of 64 Mc./sec. Each unit consists of four motor vehicles, one containing the vision transmitter (which gives a peak power of 1 kW), one the scanning and monitoring apparatus, one the power supply equipment, and one carrying the fire-escape ladder which supports the transmitting aerial. Special balanced-pair television cables have been installed by the Post Office between Broadcasting House and Alexandra Palace and also between many of the principal centres in London and Broadcasting House. Where this cable is available the radio link is unnecessary and only the scanning van and possibly the power supply van are required.

The range of the special cable can be extended by the use of the third available method, which consists of using ordinary pairs in Post Office telephone cables. Recent work has shown that by careful equalization of the frequency characteristic such pairs can be successfully used to send vision signals over a maximum distance of 4 miles. In this case it is necessary to use special terminal equipment and two intermediate line repeaters, with equalizers, installed at Post Office exchanges along the route. It is, however, unnecessary to interfere with the lines themselves.

Television Overseas

In the U.S.A. a great deal of experimental work has been done, but technical and financial problems still delay the opening of a regular service for home viewers. It is intended to demonstrate television to the public at the New York World Fair in April, 1939, and both the Columbia Broadcasting System and the National Broadcasting Co. will be transmitting television programmes for public reception, the latter from the transmitter situated in the Empire State Building and the former from a new transmitter which is being erected in the Chrysler Building. These vision transmitters work on frequencies of the order of 50 Mc./sec. (6 m.) and deliver a peak output power of 25 to 30 kW to the aerial.

In Germany a 180-line television service has been operated in Berlin for several years, but receivers were not available to the public for home use, viewing facilities being provided in public viewing rooms. A 180-line television telephone service between Berlin and Leipzig has been available to the public since March, 1936, with an extension to Munich in the summer of 1938, but it was only at the beginning of November, 1938, that a public television service, using 441 lines interlaced, 25 pictures per second, was opened on a regular schedule in Berlin and receivers made available for home reception.

In France public demonstrations of television have been carried out from an experimental station at the Eiffel Tower for several years, but high-definition transmissions did not start until the Paris Exhibition in 1937. The new high-definition transmitter has a peak power of 30 kW and works on a frequency of 46 Mc./sec. (6·53 m.). Regular transmissions are now made and receivers are available for home reception. Four different systems of transmission are being used experimentally, no one system having been selected as a standard at the present time (December, 1938).

Experimental work is being done in a number of other countries, notably in Holland where an experiment has been made using 567 lines interlaced at 25 pictures per second.

There is one aspect of television development which must be recorded with regret, and that is that there seems no likelihood of international standardization with regard to definition and picture frequency standards. It is true that no serious attempt has been made so far to bring this about, but the unfortunate fact remains that the standards at the moment are different in the following countries:—

	Lines per picture	Pictures per second	Modulation
America . .	441	30	Negative
Germany . .	441	25	Positive
Great Britain . .	405	25	Positive
France	Various between 440 and 455	25	Positive

It may be that in the distant future international television cables may become available, and then the absence of a common standard for European countries might become a serious matter.

DISCUSSION ON "MULTIPLE REFLECTIONS BETWEEN TWO TUNED RECEIVING ANTENNAE"*

Dr. G. H. Brown (U.S.A.) (*communicated*): The authors have presented a treatment of a tuned receiving antenna and a tuned parasitic reflector. I believe they have made the problem somewhat complex. Before taking up that aspect of the presentation, however, I wish to comment on their introductory statements. It appears that they have somewhat misinterpreted the previous work. On page 424 (vol. 83) they quote from my 1937 paper in the *Proceedings of the Institute of Radio Engineers*, but the statement attributed to me had no connection with the case of a receiving antenna with a parasitic reflector. Actually, in that particular section of my paper, I was treating the case of a single transmitting antenna (antenna 0) and a single receiving antenna (antenna 1). Just ahead of my statement concerning the weak current in antenna 1, I stated that the antennae were separated a great distance, "many miles if we wish."

Following this treatment, I did attack the problem of a receiving antenna with a parasitic reflector, taking account of the to-and-fro reflections by the method of mutual impedances. For one specific case—that of two tuned receiving antennae placed broadside to the impinging field—I obtained equation (103) in my paper, which may be rewritten thus:—

$$\frac{\bar{I}_0}{\bar{I}'_0} = \frac{1}{1 + \dfrac{|Z_{01}|}{R_{00}} \cos\theta_m + j\dfrac{|Z_{01}|}{R_{00}} \sin\theta_m} \quad . \quad . \quad (a)$$

where θ_m is the phase angle of the mutual impedance, Z_{01}, and j is the operator ordinarily used in alternating-current notation. The absolute value of equation (*a*) is

$$\left|\frac{I_0}{I'_0}\right| = \frac{1}{\sqrt{\left[\left(1 + \dfrac{|Z_{01}|}{R_{00}}\cos\theta_m\right)^2 + \left(\dfrac{|Z_{01}|}{R_{00}}\sin\theta_m\right)^2\right]}} \quad (b)$$

or

$$\left|\frac{I_0}{I'_0}\right| = \left[1 + \frac{2|Z_{01}|\cos\theta_m}{R_{00}} + \frac{|Z_{01}|^2}{R_{00}^2}\right]^{-\frac{1}{2}} \quad . \quad . \quad (c)$$

Equation (*c*) bears a striking resemblance to the authors' equation (2) when due consideration is given to sign.

Let us now turn to the circuit shown in Fig. A, which need not be considered as an antenna circuit, but any simple network. It is simply a coupled circuit, with the primary circuit impedance Z_{00}, a secondary circuit impedance Z_{11}, and a mutual impedance Z_m between circuits.

If the secondary circuit were opened or removed, the current in the primary due to the voltage E_0 is

$$\bar{I}'_0 = \bar{E}_0/\bar{Z}_{00} \quad . \quad . \quad . \quad . \quad . \quad (d)$$

But, if the circuits are coupled together, this component of current will induce a current in the secondary which will be

$$\bar{I}'_1 = -\bar{I}'_0\bar{Z}_m/\bar{Z}_{11} \quad . \quad . \quad . \quad . \quad . \quad (e)$$

This component of current will induce a further component in the primary which will be

$$\bar{I}''_0 = -\bar{I}'_1\bar{Z}_m/\bar{Z}_{00} = \bar{I}'_0\bar{Z}^2_m/\bar{Z}_{00}\bar{Z}_{11} \quad . \quad . \quad (f)$$

This component of current will induce current in the secondary, thus

$$\bar{I}''_1 = -\bar{I}''_0\bar{Z}_m/\bar{Z}_{11} = -\bar{I}'_0\bar{Z}^3_m/\bar{Z}_{00}\bar{Z}^2_{11} \quad . \quad . \quad (g)$$

This component will in turn produce a current in the primary, which will be

$$I'''_0 = -\bar{I}''_1\bar{Z}_m/\bar{Z}_{00} = \bar{I}'_0\bar{Z}^4_m/\bar{Z}^2_{00}\bar{Z}^2_{11} \quad . \quad . \quad (h)$$

We could keep on along this line indefinitely, securing a second-order term, a third-order term, a fourth-order term, etc. Then to get the current in the primary circuit, we add equations (*d*), (*f*), (*h*), and so on, to secure

$$\bar{I}_0 = \bar{I}'_0\left[1 + \frac{\bar{Z}^2_m}{\bar{Z}_{00}\bar{Z}_{11}} + \frac{\bar{Z}^4_m}{\bar{Z}^2_{00}\bar{Z}^2_{11}} + \frac{\bar{Z}^6_m}{\bar{Z}^3_{00}\bar{Z}_{11}{}^3} + \ldots\right] \quad (i)$$

This is the method used by the authors in treating the coupled antennae.

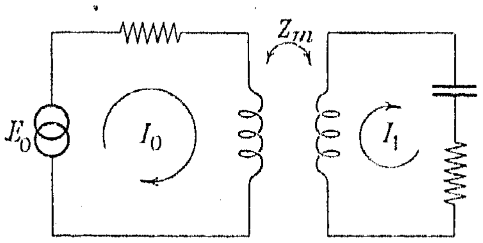

Fig. A

However, we may treat the circuit of Fig. A in another fashion. Let us write Kirchoff's law for the two circuits. Then

$$\bar{E}_0 = \bar{I}_0\bar{Z}_{00} + \bar{I}_1\bar{Z}_m \quad . \quad . \quad . \quad . \quad (j)$$

and

$$0 = \bar{I}_0\bar{Z}_m + \bar{I}_1\bar{Z}_{11} \quad . \quad . \quad . \quad . \quad (k)$$

Solving these simultaneous equations, we find

$$\bar{I}_0 = \frac{\bar{E}_0 Z_{11}}{\bar{Z}_{00}\bar{Z}_{11} - Z^2_m} \quad . \quad . \quad . \quad . \quad (l)$$

Dividing (*l*) by (*d*),

$$\frac{I_0}{I'_0} = \frac{Z_{00}Z_{11}}{Z_{00}Z_{11} - Z^2_m} \quad . \quad . \quad . \quad . \quad (m)$$

Dividing numerator and denominator of (*m*) by $Z_{00}Z_{11}$,

$$\frac{I_0}{I'_0} = \frac{1}{1 - (Z^2_m/Z_{00}Z_{11})} \quad . \quad . \quad . \quad . \quad (n)$$

That equation (*n*) is identical with (*i*) is apparent when it is remembered that

$$1/(1 - x) = 1 + x + x^2 + x^3 + x^4 + \ldots \quad . \quad (o)$$

Whether antennae with distributed constants or lumped circuits are involved, the problem is the same. The method of mutual impedances as advanced by Carter is quite adequate to take care of the " to-and-fro " effect. In principle, the method of mutual impedances is the same as the method proposed by Prof. Palmer and Messrs. Abson and Barker. I much prefer the simplicity of the method of mutual impedances.

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* Paper by Prof. L. S. PALMER and Messrs. W. ABSON and R. H. BARKER (see vol. 13, page 294). Reprinted from *Journal I.E.E.*, 1939, vol. 84, p. 749.

Prof. L. S. Palmer and **Messrs. W. Abson** and **R. H. Barker** (*in reply*): Our statement concerning the currents in Dr. Brown's antennae 0 and 1 is correct as long as the distance between the antennae is such that antenna 1 can be considered to be a parasitic antenna to antenna 0. It is, of course, true that the current in 1 will be negligibly small compared with that in 0 when the distance has been increased to such an extent that 1 can no longer be considered as a parasitic antenna to 0 but has become a distant receiving antenna. Dr. Brown's statement correctly referred to the latter condition, whilst ours referred to the former, and thus we inadvertently gave a wrong impression in our quotation from Dr. Brown's paper.

Concerning our method of treatment, although we feel that Dr. Brown's method of mutual impedances is more elegant mathematically, it is not very much shorter or simpler than our method of multiple reflections when applied to the general case of two antennae in a plane at any angle θ to the direction of wave propagation. This is shown below.

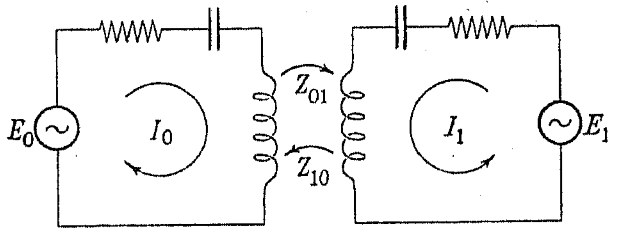

Fig. B

On the other hand, the physical picture resulting from our method of multiple reflections enables one to visualize that the approximations which we both make depend on the number of to-and-fro reflections which are taken into account in order to attain a required degree of accuracy. Dr. Brown's method does not readily lend itself to varying degrees of approximation. Consequently, if the required accuracy can be obtained by considering only one reflection (as when comparing calculated and measured critical antennae spacings) then our method is the simpler. Dr. Brown's method of mutual impedances, when applied to the general case of two similar tuned receiving antennae, is as follows, and may be compared with that given in our paper.

Two antennae, each receiving radiation from a distant transmitter, and each acting as a parasitic reflector to the other, can be represented by the circuit shown in Fig. B, where E_0 and E_1 are the instantaneous potentials due to the transmitter. For parallel radiation the magnitudes $|E_0|$ and $|E_1|$ are equal and E_1 lags behind E_0 by $\alpha = (2\pi D \cos\theta)/\lambda$, where D is the distance between the antennae, and θ is the angle between the plane containing the antennae and the direction of wave propagation.

Z_{00} = impedance of antenna 0.
I_0 = instantaneous value of the current in 0 when 1 present.
I_0' = instantaneous value of the current in 0 when 1 absent.
$= E_0/Z_{00}$
Z_{01} = impedance of radiation coupling between 0 and 1.
Z_{11}, I_1, I_1', Z_{10}, are the corresponding quantities for antenna 1.

Kirchoff's laws for the circuit are then:—

$$E_0 = I_0'Z_{00} = I_0Z_{00} + I_1Z_{01}$$
$$E_1 = I_1'Z_{11} = I_1Z_{11} + I_0Z_{10}$$

Eliminating I_1 from these equations and remembering that $Z_{01} = Z_{10}$, we get

$$I_0Z_{00}Z_{11} - I_0Z_{01}^2 = I_0'Z_{00}Z_{11} - I_1'Z_{11}Z_{01} \quad . \quad . \quad (p)$$

For two identical tuned antennae

$$Z_{00} = Z_{11} = R_{00} + jX_{00} = R_{00}$$

where R_{00} = resistance of 0, and X_{00} = reactance of 0 for the wavelength under consideration.

Also let $$Z_{01} = R_{01} + jX_{01} \quad . \quad . \quad . \quad . \quad (q)$$

Furthermore

$$I_0' = |I_0'| \sin \omega t, \quad I_1' = |I_1'| \sin(\omega t - \alpha)$$

and $$|I_0'| = |I_1'|.$$

Hence (p) becomes

$$I_0(R_{00}^2 - Z_{01}^2) = R_{00}^2|I_0'| \sin \omega t - Z_{01}R_{00}|I_0'| \sin(\omega t - \alpha)$$

or $$I_0 = \frac{|I_0'| \sin \omega t}{[1 - (Z_{01}^2/R_{00}^2)]} - \frac{Z_{01}|I_0'| \sin(\omega t - \alpha)}{R_{00}[1 - (Z_{01}^2/R_{00}^2)]}$$

Using (q) we get

$$\frac{|I_0'| \sin \omega t}{\left[\left(1 - \dfrac{R_{01}^2 - X_{01}^2}{R_{00}^2}\right) - 2j\dfrac{R_{01}X_{01}}{R_{00}^2}\right]}$$
$$- \frac{(R_{01} + jX_{01})|I_0'| \sin(\omega t - \alpha)}{R_{00}\left[\left(1 - \dfrac{R_{01}^2 - X_{01}^2}{R_{00}^2}\right) - 2j\dfrac{R_{01}X_{01}}{R_{00}^2}\right]}$$
$$= |I_0'|\left[\frac{1}{\left(1 - \dfrac{R_{01}^2 - X_{01}^2}{R_{00}^2}\right)^2 + \dfrac{4R_{01}^2X_{01}^2}{R_{00}^4}}\right]^{\frac{1}{2}} \sin(\omega t - \phi)$$
$$- \frac{|I_0'|}{R_{00}}\left[\frac{R_{01}^2 + X_{01}^2}{\left(1 - \dfrac{R_{01}^2 - X_{01}^2}{R_{00}^2}\right)^2 + \dfrac{4R_{01}^2X_{01}^2}{R_{00}^4}}\right]^{\frac{1}{2}}$$
$$\sin(\omega t - \alpha - \phi + \arctan X_{01}/R_{01}) \quad (r)$$

where ϕ is the phase lag due to the denominator $(1 - Z_{01}^2/R_{00}^2)$ operating on a sine function,

$\arctan X_{01}/R_{01} = -\beta$, $(R_{01}^2 + X_{01}^2)^{\frac{1}{2}}$ is our factor A, and $1/R_{00}$ is our factor B.

Hence $$A \cos\beta = \frac{(R_{01}^2 + X_{01}^2)^{\frac{1}{2}}}{(1 + X_{01}^2/R_{01}^2)^{\frac{1}{2}}} = R_{01}$$

Thus, on neglecting $(AB)^3$ and higher powers, (r) reduces to

$$I_0 \simeq [1 + 2(AB)^2 \cos 2\beta]^{\frac{1}{2}}$$
$$[|I_0'| \sin(\omega t - \phi) - AB|I_0'| \sin(\omega t - \phi - \alpha - \beta)]$$

which, on subtracting vectorially, becomes

$$\left|\frac{I_0}{I_0'}\right| \simeq \left[1+2(AB)^2\cos 2\beta\right]^{\frac{1}{2}}\left[1+(AB)^2+2(AB)\cos(\alpha+\beta)\right]^{\frac{1}{2}}$$

$$\simeq \left[1+2(AB)\cos(\alpha+\beta)+(AB)^2(1+2\cos 2\beta)\right]^{\frac{1}{2}}$$

We thus recover our general equation (4*a*) using Dr. Brown's method of mutual impedances.

It is of interest, and we believe of value, to have had this opportunity of discussing and comparing the two methods of attacking the problem.

MEETINGS OF THE WIRELESS SECTION

147th MEETING OF THE WIRELESS SECTION, 7th DECEMBER, 1938

Mr. A. J. Gill, B.Sc.(Eng.), Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting held on the 2nd November, 1938, were taken as read and were confirmed and signed.

The following papers were read and discussed: " Long Feeders for Transmitting Wide Side-Bands, with reference to the Alexandra Palace Aerial-Feeder System " (see page 111), by Messrs. E. C. Cork, B.Sc.(Eng.), Associate Member, and J. L. Pawsey, Ph.D.; and " E.M.I. Cathode-Ray Television Transmission Tubes " (see page 131), by Messrs. J. D. McGee, M.Sc., Ph.D., and H. G. Lubszynski, Dr.Ing.

A vote of thanks to the authors, moved by the Chairman, was carried with acclamation.

148th MEETING OF THE WIRELESS SECTION, 4th JANUARY, 1939

Mr. A. J. Gill, B.Sc.(Eng.), Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting held on the 7th December, 1938, were taken as read and were confirmed and signed.

Sir Noel Ashbridge, B.Sc.(Eng.), then delivered a lecture entitled " Broadcasting and Television."

A vote of thanks to the lecturer, moved by the Chairman, was carried with acclamation.

15th INFORMAL MEETING OF THE WIRELESS SECTION, 24th JANUARY, 1939

Mr. A. J. Gill, B.Sc.(Eng.), Chairman of the Section, took the chair at 6.30 p.m.

The minutes of the Informal Meeting held on the 22nd November, 1938, were taken as read and were confirmed and signed.

A discussion, opened by Mr. M. V. Callendar, took place on " Automatic Tuning Devices."

At the conclusion of the discussion a vote of thanks was accorded to Mr. Callendar for his introductory remarks.

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